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# INTERFERENCE REDUCTION GUIDE

## FOR DESIGN ENGINEERS

**VOLUME 1**

**CHAPTER 1. INTRODUCTION TO INTERFERENCE REDUCTION DESIGN PRACTICE**
**CHAPTER 2. GROUNDING, BONDING AND SHIELDING DESIGN THEORY AND PRACTICE**



*U.S. ARMY ELECTRONICS LABORATORIES*
Fort Monmouth, N. J.




*Prepared by:*
**FILTRON COMPANY, INC., NEW YORK**
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# FOR

# DESIGN ENGINEERS

VOLUME I

1 August 1964



ELECTROMAGNETIC ENVIRONMENT DIVISION
U.S. ARMY ELECTRONICS LABORATORIES
FORT MONMOUTH, NEW JERSEY

## INTERFERENCE REDUCTION GUIDE FOR DESIGN ENGINEERS

### PREFACE

Interference generated by electrical and electronic equipment can seriously interfere with the generation, transmission, and reception of desired signals by communications-electronic equipment. Such interference problems are on the increase because of the growing reliance upon communications-electronic equipment in military operations, increasing transmitter powers, better receiver sensitivities, and the crowding of the electromagnetic spectrum in which such equipment operates. Interference, that results from spurious responses or spurious radiations in electronic equipment, detracts from the usefulness of such equipment, reduces its efficiency, or may even make it completely inoperable.

In designing equipment, the engineer must consider how the functional requirements of the equipment are affected by interference characteristics. In this way, interference problems can be minimized by maximum utilization of interference-free circuits and components and application of other interference control techniques.

Interference-proof design can be very, very expensive if the original design does not include interference reduction. It has been the usual practice to design communications-electronic equipment with very little thought given to the adverse electromagnetic environment in which the equipment must operate. This has been caused by several things, among which are:

(a) Interference reduction is often left out of the initial design in the expectation that if trouble is encountered later, "field fixes" will take care of the problem (as in many cases they do)

(b) Many engineers believe that filtering and shielding alone do an adequate job

(c) There has been a tendency to believe that interference specifications are needlessly harsh, or cannot be complied with even when the requirements are recognized as legitimate

(d) There are perhaps a few otherwise very good design engineers who just don't know how to incorporate interference reduction into their basic designs

(e) Time and money pressures sometimes lead to the erroneous belief that interference reduction at the design stage is too expensive

This guide is an attempt to provide the engineer with the necessary background and techniques to enable him to minimize the interference generation and interference susceptibility of the equipment that he designs. The guide is not intended for the interference experts. It is not a manual for field fixes, nor for interference suppression after field tests have shown interference problems. It is intended for the use of design engineers with little or no interference reduction experience. Some tables, curves and other handbook-type data are included but only to the extent that such data is not readily available in standard handbooks or manuals.

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# CHAPTER 1

# INTRODUCTION TO INTERFERENCE REDUCTION DESIGN PRACTICE

## Section I. DEFINITIONS

### 1-1. General

The definitions presented are, for the most part, peculiar to interference engineering. They are defined as they relate specifically to the study of interference control for the U.S. Army.

### 1-2. Ambient Interference

Ambient interference is that level of interference indicated by the interference measuring set when sited, set up, tuned, and calibrated for the intended interference measurement, with the equipment under test turned off.

### 1-3. Bond

A bond is an electrical interconnection made with a low-resistance material between such items as chassis, metal shield cans, panels, cable shielding braid, and other equipotential points to eliminate undesirable interaction resulting from high-impedance paths between them.

### 1-4. Broadband Interference

Broadband interference includes impulse noise, thermal noise, shot noise, and other nonsinusoidal interference whose energy is distributed over a spectrum of frequencies that is wide when compared with the bandwidth of the interference measuring equipment. The response of an interference measuring set to broadband interference is a function of the effective bandwidth of the receiver.

### 1-5. Cross-Modulation and Crosstalk

Cross-modulation is the modulation of a desired signal by an undesired signal. Crosstalk is interference caused by energy being coupled from one circuit to another by stray coupling.

### 1-6. Decoupling

Decoupling is the prevention of transfer of interference energy from one circuit to another.

1-7. Impedance Stabilization Network

An impedance stabilization network provides a standard line impedance across which conducted interference is measured.

1-8. Impulse Noise

Impulse noise is noise or interference due to electrical disturbances having abrupt changes of short duration. Interference from impulse noises comprises a series of nonoverlapping transient signals.

1-9. Insertion Loss

Insertion loss is the total reduction in power delivered to a load or detector that results from insertion of an electrical network, element, or shield between the load and the power source. Insertion loss is the sum of reflection losses and attenuation.

1-10. Integral Suppression

An integrally suppressed electrical or electronic subassembly is an item that was manufactured with the necessary shielding and suppression components as a design requirement.

1-11. Interference

Interference is any conducted, induced, or radiated electrical disturbance, including transients, that can cause undesirable responses or otherwise impair the operation of electrical or electronic equipment.

1-12. Interference Tests

Interference tests are radiated, conducted, and/or susceptibility tests performed under controlled conditions, preferably in a low-ambient noise area, to determine specification and/or design compliance.

1-13. Narrowband Interference

Narrowband interference comprises cw signals, modulated or unmodulated, such as a carrier frequency, or signals whose energy is distributed over a spectrum of frequencies that is narrow when compared with the bandwidth of the receiver or interference measuring equipment.

## 1-14. Shield

A shield is a conducting enclosure surrounding a source of interference or a susceptible circuit that is designed to reduce the radiation of interference or to prevent a susceptible circuit from being affected by interfering signals.

## 1-15. Spurious Emission and Response

Emission of electromagnetic energy at any frequency or frequencies other than at the designed operating frequency is considered spurious emission. A response of a receiver to any frequency or frequencies other than the one at which it is adjusted and designed to operate is considered spurious response.

## 1-16. Susceptibility

Susceptibility is the response of a piece of equipment or component to any signal applied at other than the normal input terminals, or to any signal other than the desired one applied at the normal input terminals, relative to its response to a desired signal applied at the normal input terminals.

## 1-17. Wave Trap

A wave trap is a resonant circuit usually connected into an antenna system to suppress undesired signals at a discrete frequency.

## Section II. CHARACTERISTICS OF INTERFERENCE

### 1-18. General

Interference signals are grouped into narrowband or broadband types (fig. 1-1). Broadband interference is further divided into random and impulse interference. Random interference consists of closely spaced electromagnetic impulses that are not clearly distinguishable from one another. The impulses are frequent and overlap, with sharp peaks exceeding the average level. Impulse interference is characterized by sharp pulses that are relatively infrequent and clearly separated; for example, thermal agitation and atmospheric interference. Impulse interference may be generated by an internal combustion engine ignition system, power line discharges, motor brush sparking, electronic equipment, or other electrical or electromechanical devices.

### 1-19. Broadband Interference Generation

a. Broadband interference is generated by any steep, sharp-angled, or short-duration nonsinusoidal current or voltage waveform. The frequency range of interference generated by an interference source depends on the shape of the wave produced. Any irregularly shaped wave or pulse is composed of a fundamental sine wave, to which are added various harmonics of different frequencies and amplitudes having definite phase relationships to the fundamental. The equivalence of the pulse and the group of individual waves can be shown by adding up the contributions due to the separate waves and showing that their sum is equal to the original wave.

b. The shape of a complex wave depends on the number of harmonics present, their relative amplitudes, and phase relationships. In general, the steeper the wavefront, the more harmonics it contains. Thus, a perpendicular wave front is composed of a fundamental and an infinite number of harmonics.

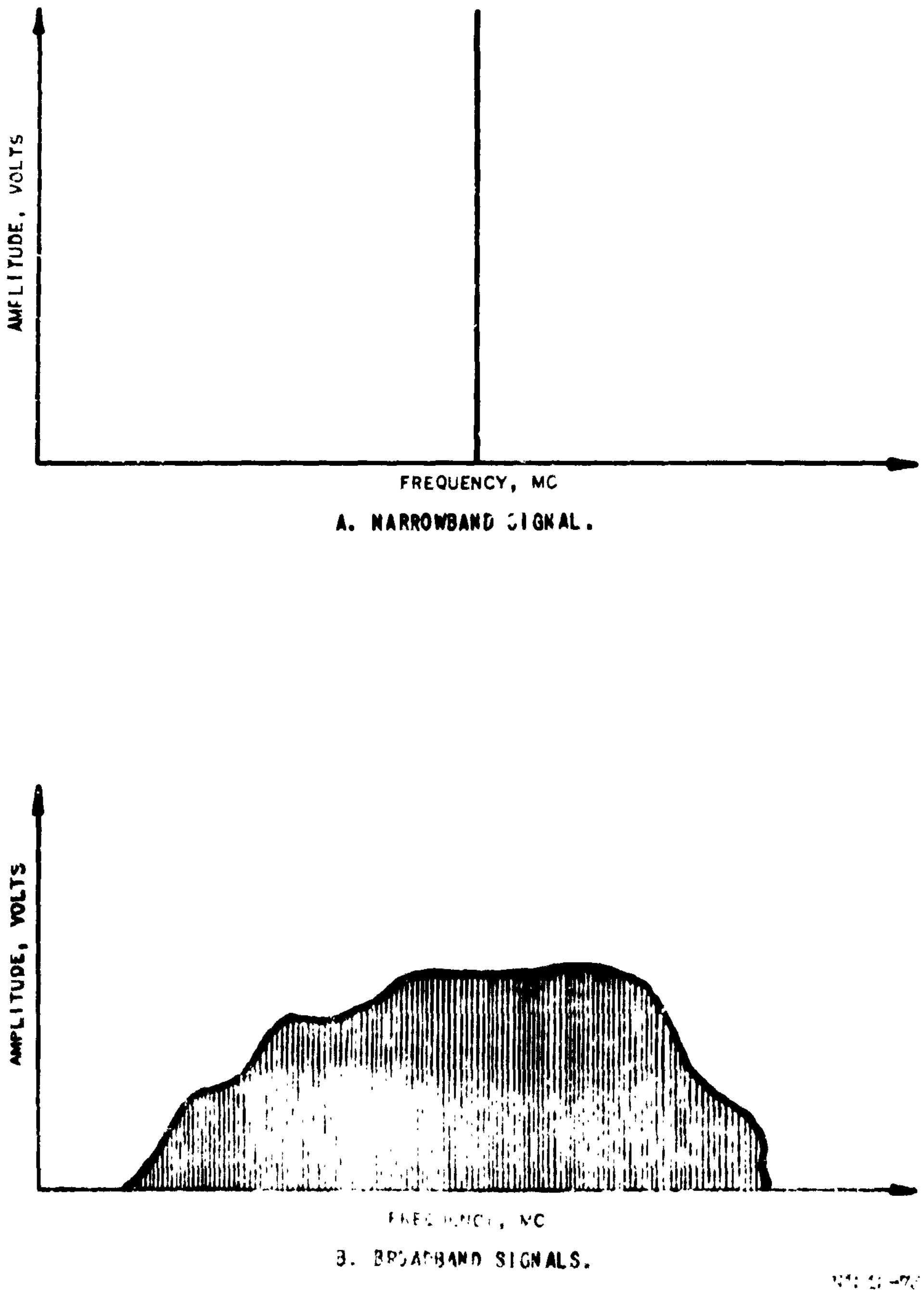



Figure 1-1. Interference Signals

c. Figure 1-2 illustrates the formation of a square wave. If a square wave is compared with a sine wave (A) of the same frequency, it can be seen that they are considerably different. However, if another sine wave (B) of smaller amplitude, but three times the frequency, is added in phase to A, the resultant wave more nearly approaches the square wave. The resultant wave is curve C. When the fifth harmonic (D) is added, the sides of the new wave are steeper than before. The new wave is the line E. Addition of the seventh harmonic (F) of even smaller amplitude makes the sides of the composite wave even steeper. Addition of more and more odd harmonics, in phase with the fundamental, brings the resulting wave nearer and nearer to the square wave.

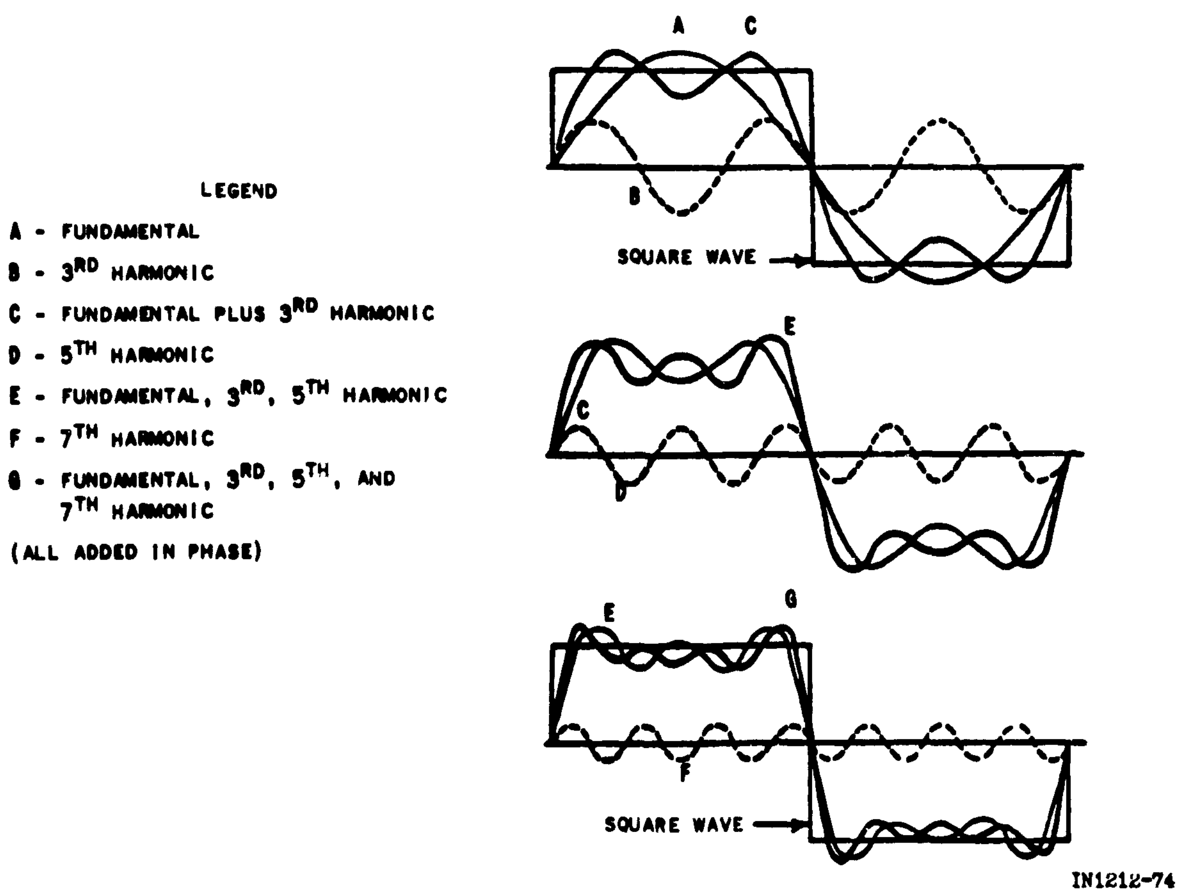



Figure 1-2. Square-Wave Formation

d. The composition of a peaked wave, which closely corresponds to transient voltage waves, is illustrated on figure 1-3. As more harmonics are added to the wave out of phase with the fundamental, it becomes more and more peaked. The lowest frequency (the fundamental) in a pulse or transient has the greatest amplitude. Each harmonic contained in the pulse is of smaller amplitude, except where some part of the circuit may resonate with a harmonic, thus increasing its amplitude. If an oscilloscope is connected across the output terminals of an interference source, it is possible (provided the frequency is not too high) to determine the general shape of the voltage wave produced by the source. The wave shape will alter appreciably if the load and line impedances are changed. Since the original wave shape, with no load, depends on the internal impedance of the source, it can be seen that all three impedances (load, line, and source) affect the wave shape. Because of resonant conditions, the L and C components of these impedances attenuate certain frequencies and amplify others, which may greatly increase the interference.

1-20. Theoretical Considerations of Broadband Interference Generation

a. For the repetitive rectangular pulse shown on figure 1-4, a Fourier analysis of its theoretical spectral content is shown (fig. 1-5). By doubling the pulse repetition rate with a given pulse width, the spacing of successive harmonics is doubled, and the amplitude of a given order harmonic is increased (fig. 1-6).

b. Broadband interference level is inversely proportional to pulse width, directly proportional to steepness of the edges of a pulse, and determined to a great extent by pulse width rather than by pulse shape (provided the pulse approximates a square, triangle, trapezoid, or sine wave, as shown on figure 1-7). The formula for determining the interference level of a rectangular pulse is shown on figure 1-8. Using this formula, the interference level of a 1 volt, 1 μsec rectangular pulse is plotted showing the contribution of each significant factor. A line drawn tangent to the peaks of the lobes of the envelope of the spectral lines is shown to approximate the interference level closely. Figure 1-9 is a similar straight-line approximation of the interference level of a 1 volt, 1 μsec trapezoidal pulse, having a rise time of 0.1 μsec. Formulas for calculating the interference level for any trapezoidal pulse as a function of pulse peak amplitude, pulse duration, and pulse rise time are also given on figure 1-9.

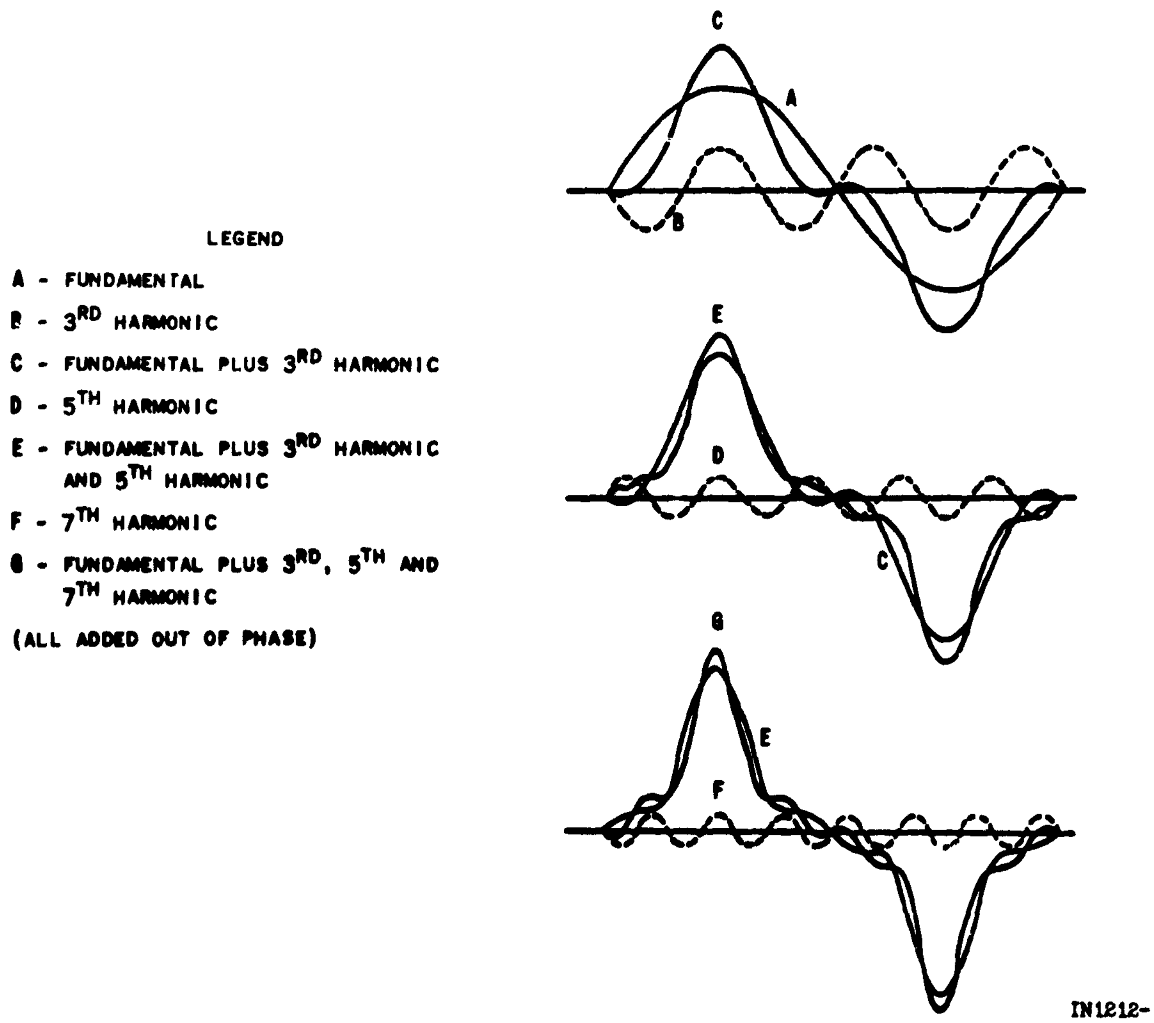
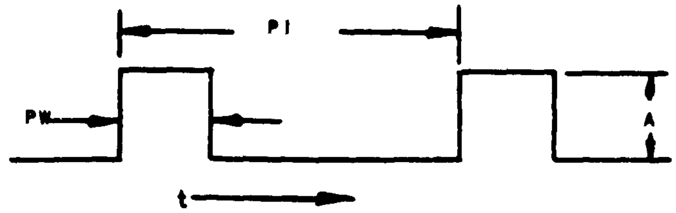



Figure 1-3. Harmonics Contained in a Peak Wave




where: PW = Pulse width
PI = Pulse interval = $\frac{1}{PRR}$

$$An = 2A\frac{(PW)}{PI} \cdot \left[\frac{\sin \frac{n\pi\ (PW)}{PI}}{\frac{n\pi\ (PW)}{PI}}\right]$$

where: An = amplitude of the $n^{th}$ harmonic
n = 0, 1, 2, 3, 4, . . .

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Figure 1-4. Fourier Analysis of a Repetitive Rectangular Pulse

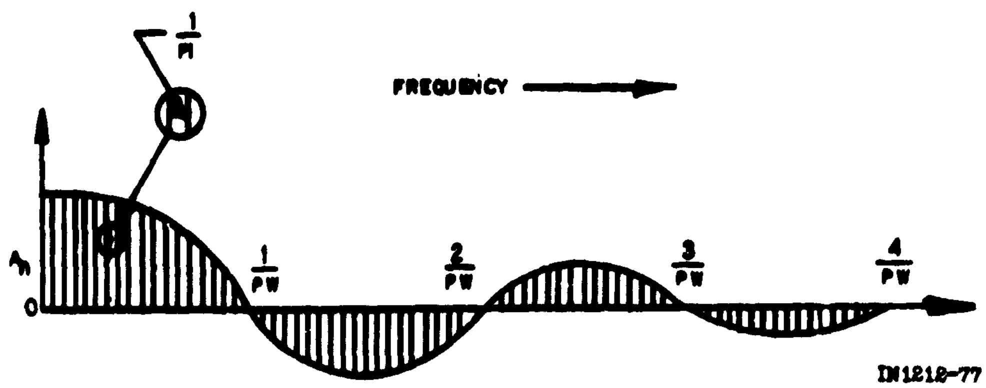


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Figure 1-5. A Plot of Harmonic Frequency Versus Harmonic Amplitude for the Pulse Shown on Figure 1-4

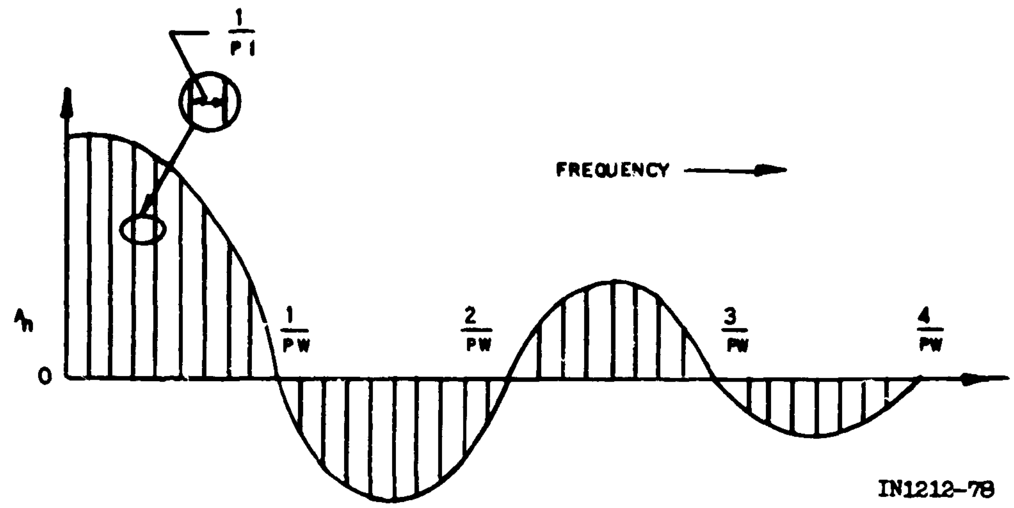


Figure 1-6. Doubling the Pulse Repetition Rate of Figure 1-5 Plot (Spacing Between Harmonics Doubled and Amplitude of Harmonics Increased)

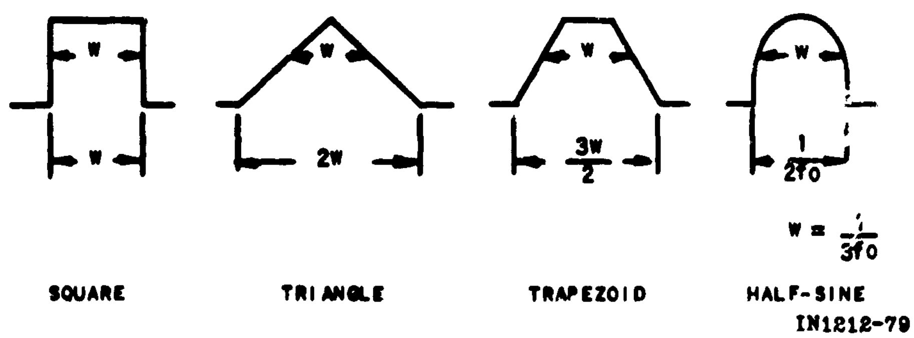


Figure 1-7. Amplitude Curves For Various Waveforms

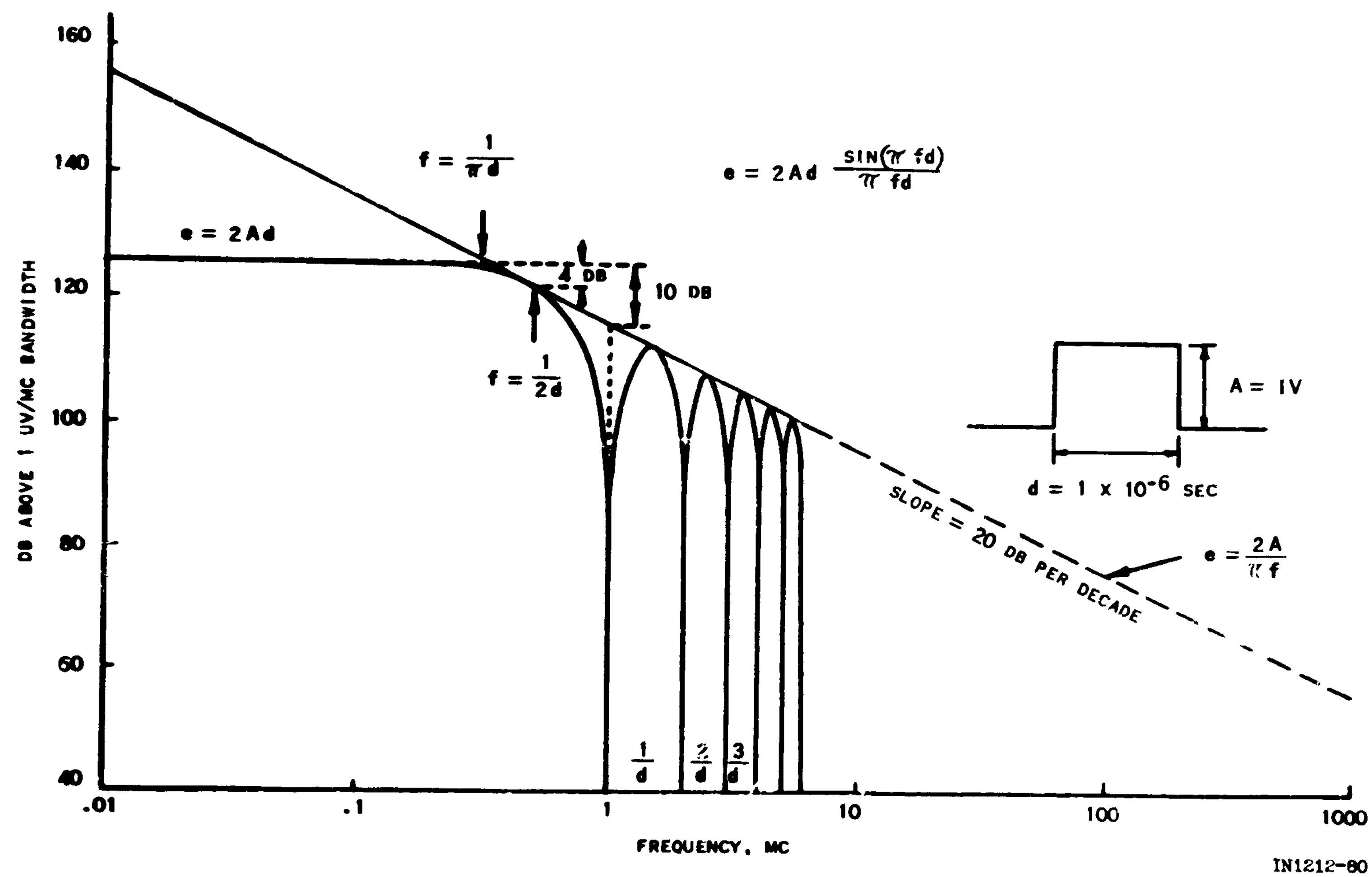


Figure 1-8. Interference Level for a 1 Volt, 1 μsec Rectangular Pulse

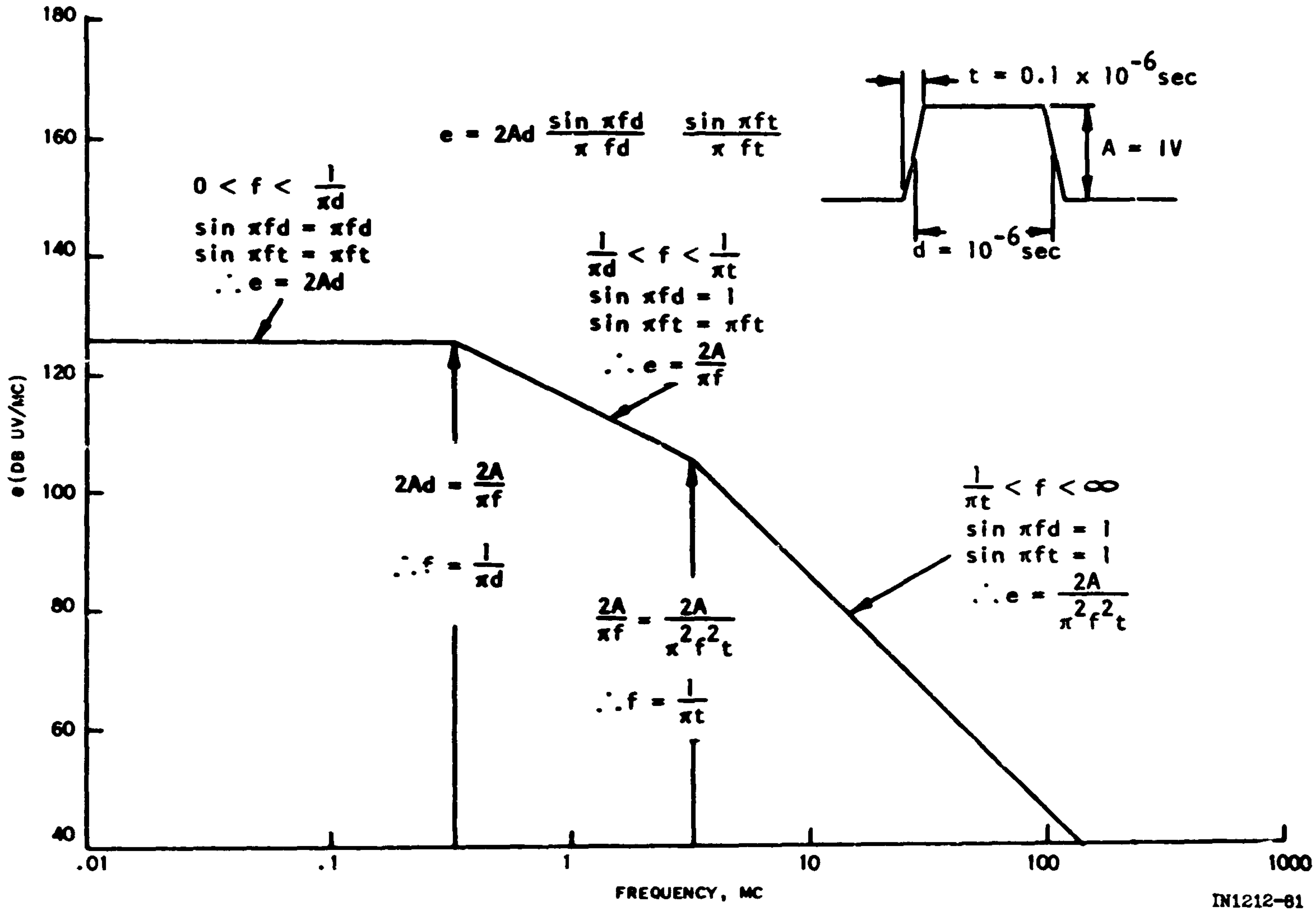


Figure 1-9. Interference Level for a 1 Volt, 1 μsec Trapezoidal Pulse

c. Envelopes of interference levels for one-volt trapezoidal pulses are plotted on figure 1-10. The interference level at low frequencies is determined only by pulse duration; at high frequencies, only by the rise time of the pulse. The interference levels for any trapezoidal pulse are given on figure 1-11. The area under the pulse, maximum amplitude of the pulse, and rate of rise of the pulse define the envelopes of interference in the three frequency regions. The average pulse duration and the pulse rise time determine the corner frequencies.

d. The interference levels for eight common pulse shapes are shown on figure 1-12. All have a one-volt peak amplitude and an average pulse duration of 1 μsec. Because all of the pulses have the same area, the interference levels at low frequencies are identical. At frequencies less than $\frac{1}{d}$, the interference level equals 2Ad or 126 db above a μv per mc. Above a frequency of approximately $\frac{1}{d}$, the interference level drops at a rate determined by the shape of the rise and fall time of the pulse. For a rectangular pulse or a clipped sawtooth pulse (both of which have step functions), the spectrum extends to higher frequencies; it decreases as the first power of frequency, or at a 20-db-per-decade rate. For a trapezoidal pulse, a critically damped exponential pulse, a triangular pulse, and a cosine pulse (all of which have sharp corners), interference decreases as the second power of frequency, or at a 40-db-per-decade rate. For a cosine-squared pulse, which has small corners, the interference decreases as the third power of frequency, or at a 60-db-per-decade rate. For a gaussian pulse, the smoothest pulse considered, the interference drops at a rate that increases with frequency.

e. The curves on figure 1-13 are identical to those on figure 1-12, but the ordinate and abscissa scales have been changed so that the curves can be applied to a pulse of any voltage amplitude and any pulse duration. The ordinate gives the number of decibels below 2Ad, and the abscissa is plotted in frequency in terms of $\frac{1}{d}$. To find the interference level of any pulse, calculate 2Ad from the peak pulse amplitude and the average pulse duration and select the curve which most nearly resembles the given pulse shape. This curve will give the number of decibels below 2Ad for any frequency.

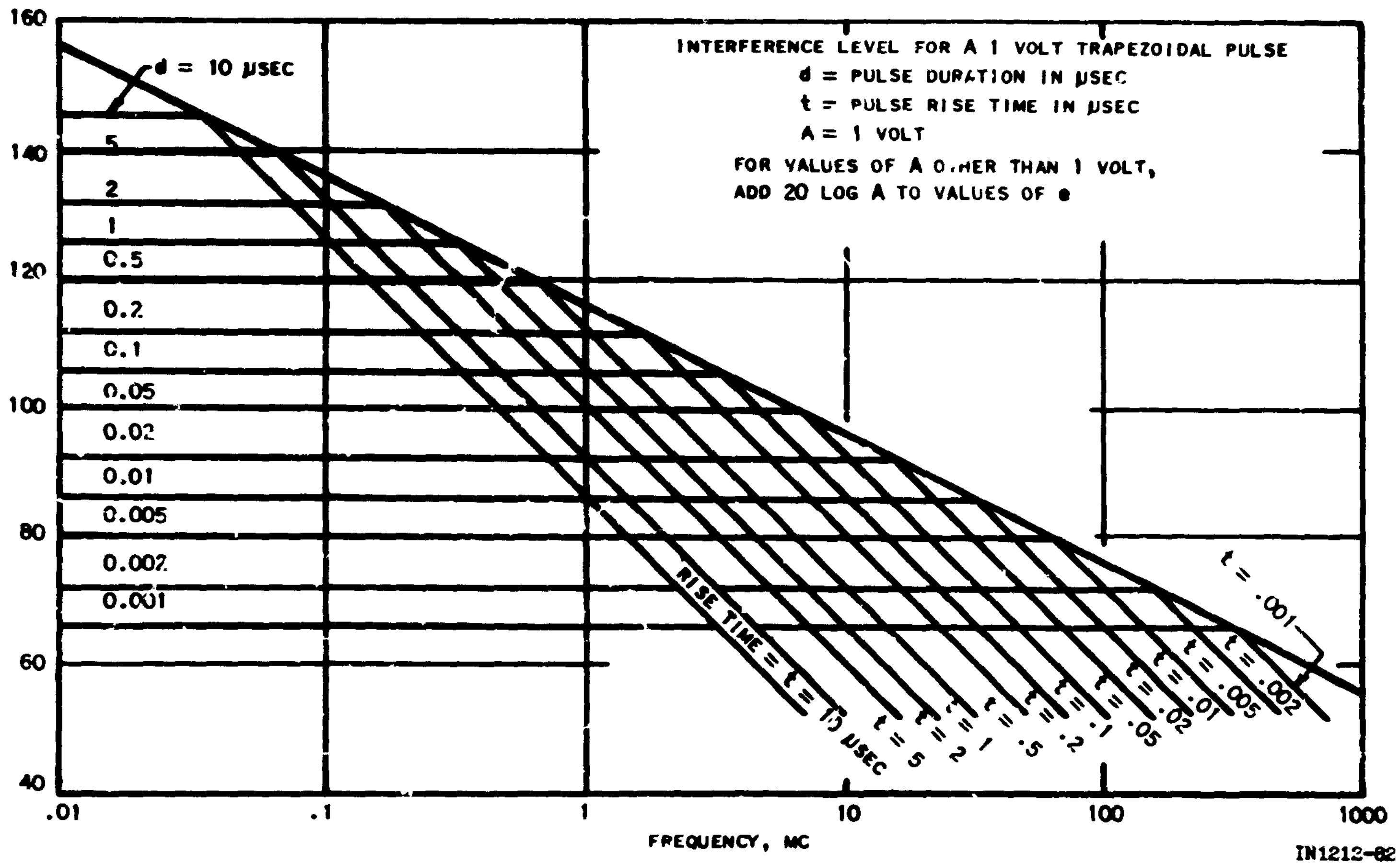


Figure 1-10. Trapezoidal Pulse Interference

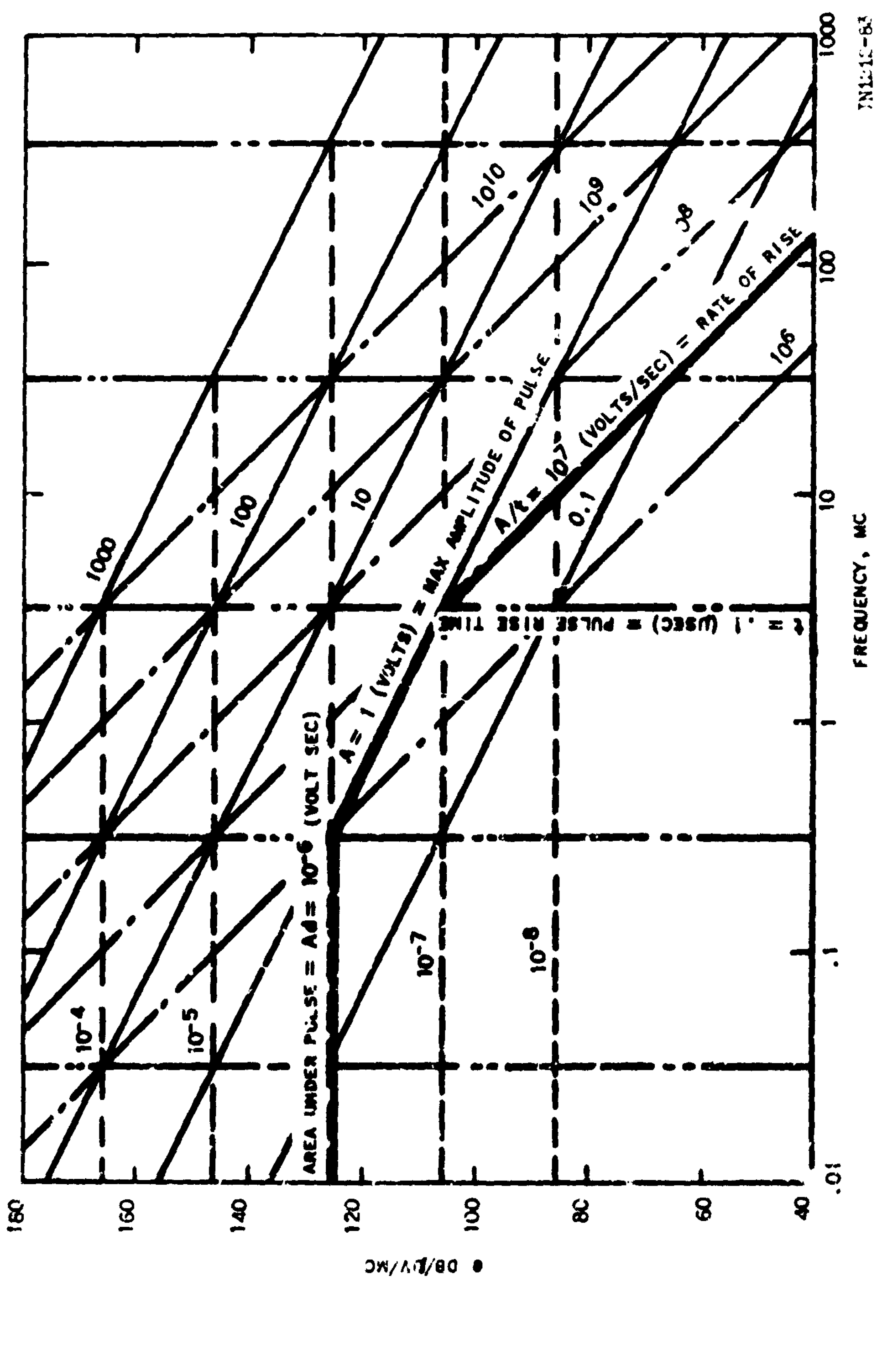


Figure 1-11. Interference Levels for Trapezoidal Pulses

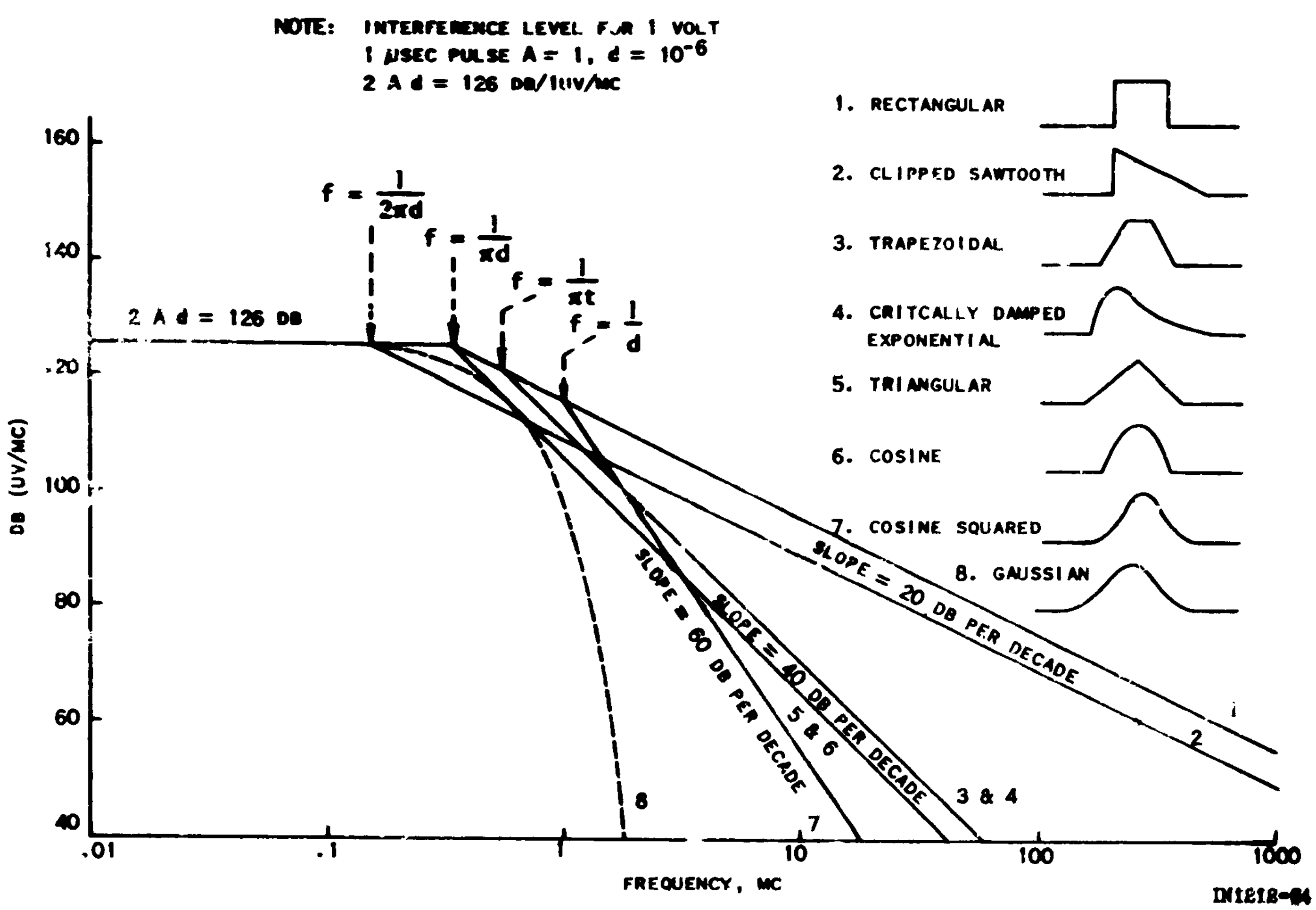


Figure 1-12. Interference Levels for Eight Common Pulses

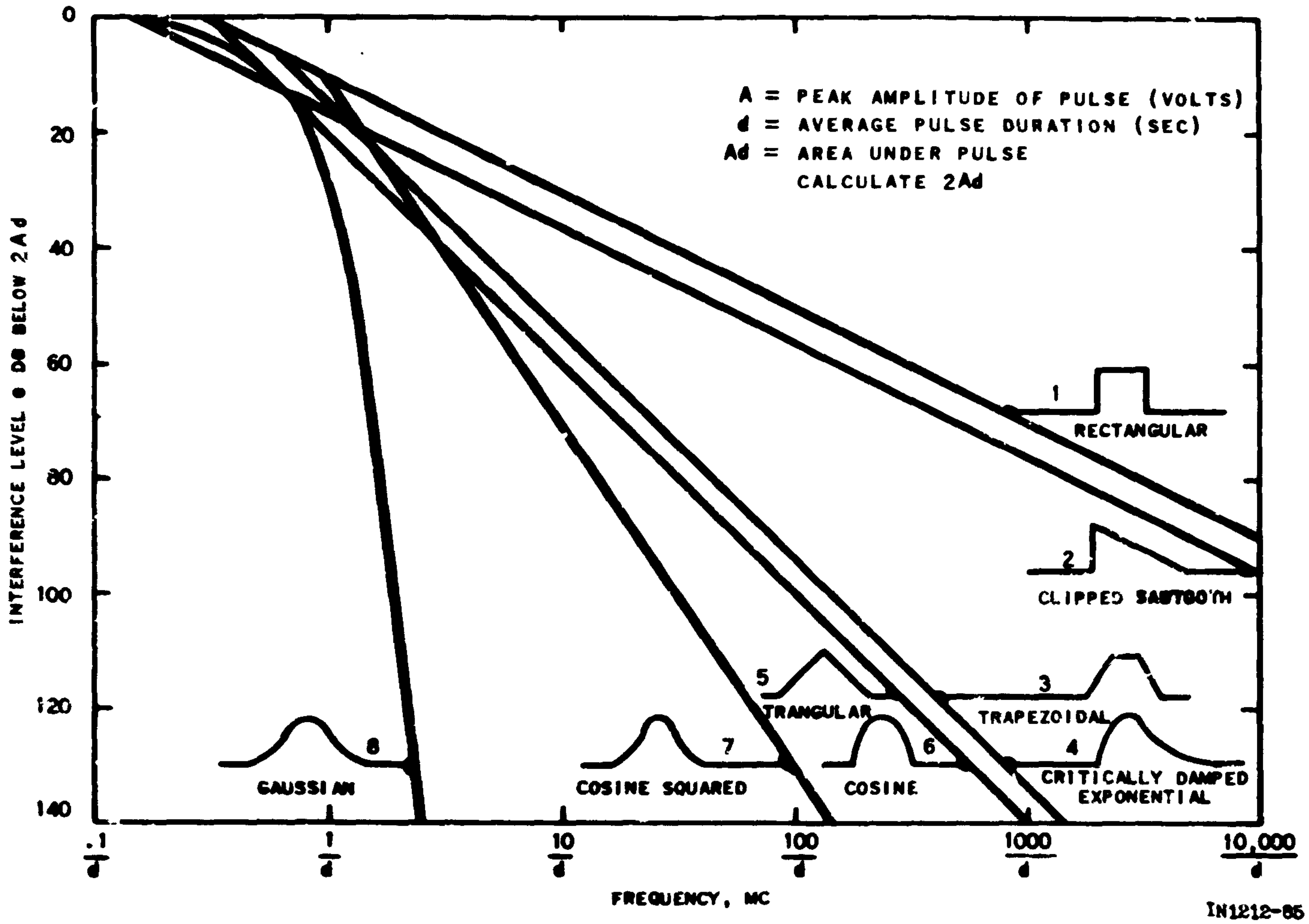


Figure 1-13. Interference Levels for Various Pulse Shapes

f. Although the pulse shapes analyzed here are ideal, the same methods may be used for any transient by considering the transient to be made up of a number of rectangular and triangular pulses. For any pulse shape, the interference level at low frequencies depends only on the area under the transient; at high frequencies, the level depends on the number and steepness of the slopes. The curves on figure 1-13 are the envelopes of the frequency lobes, each lobe being an envelope of the harmonic lines that make up the transient. Although the true curves have numerous sharp nulls, the envelope is less than 3 db from the average interference level and represents the worst case of maximum interference level. From these curves the designer can predict the type and level of interference that is going to be generated and determine the amount of suppression to design into his circuitry.

1-21. Narrowband Interference Generation

There are seven principal sources of narrowband interference generation:

1) Unintentional harmonics generated by the nonlinearity of active circuit elements, such as oscillators and amplifiers, which are designed to operate with a pure sine wave

2) Intentional nonlinearities, such as in modulators, mixers, and rectifiers, which generate a group of harmonics and intermodulation products along with desired products

3) Frequencies intentionally generated for use within the equipment and not intended for emissions

4) Inadequate limitation of the bandwidth of the modulating signal transmitter

5) Nonlinear modulation (including overmodulation), and, in the case of signals, amplitude distortion after modulation

6) Extraneous modes of oscillation of magnetrons, generating unpredictable frequencies not basically related to the generation of the desired frequency

7) Radiated power greater than needed for the particular application

## 1-22. Natural Sources of Interference

Certain natural occurrences produce sufficient electromagnetic energy to interfere with normal operation of communications, navigation, and other electronic equipment. Some of the more important natural sources are:

a. Atmospheric Interference. Atmospheric interference is erratic in character. It consists of short, randomly recurring pulses completely without regularity in phase or amplitude, and its energy is not confined to a particular band. Its average power level is relatively constant during any given hour although it may vary considerably over a longer period of time. Atmospheric interference fluctuations depend on such things as frequency, time of day, season, location, and weather conditions.

(1) Atmospheric interference occurs predominately in frequency regions below 50 mc; it dominates all other interference sources below 30 mc; and it is usually the limiting factor in communications within this spectrum. Atmospheric interference decreases rapidly above 30 mc. Figure 1-14 shows typical daytime and nighttime atmospheric interference levels as measured by a ground antenna situated in a high interference location.

(2) Reception of atmospheric interference may be reduced by decreasing antenna beamwidth, sidelobe level, and response bandwidth, while increasing antenna directivity and transmitted power. The values presented on figure 1-14 are typical for tropical ground stations. Lower values will be experienced by a space vehicle several hundred miles above the earth. The ionosphere acts as a partial shield for the majority of frequencies blanketed by atmospheric noise. Most of the incident energy will be reflected by the ionosphere below the ionosphere critical frequency. Even though atmospheric interference is shielded, cosmic and solar noise will, however, very likely be present at these frequencies to an extent much greater than at the lower altitudes.

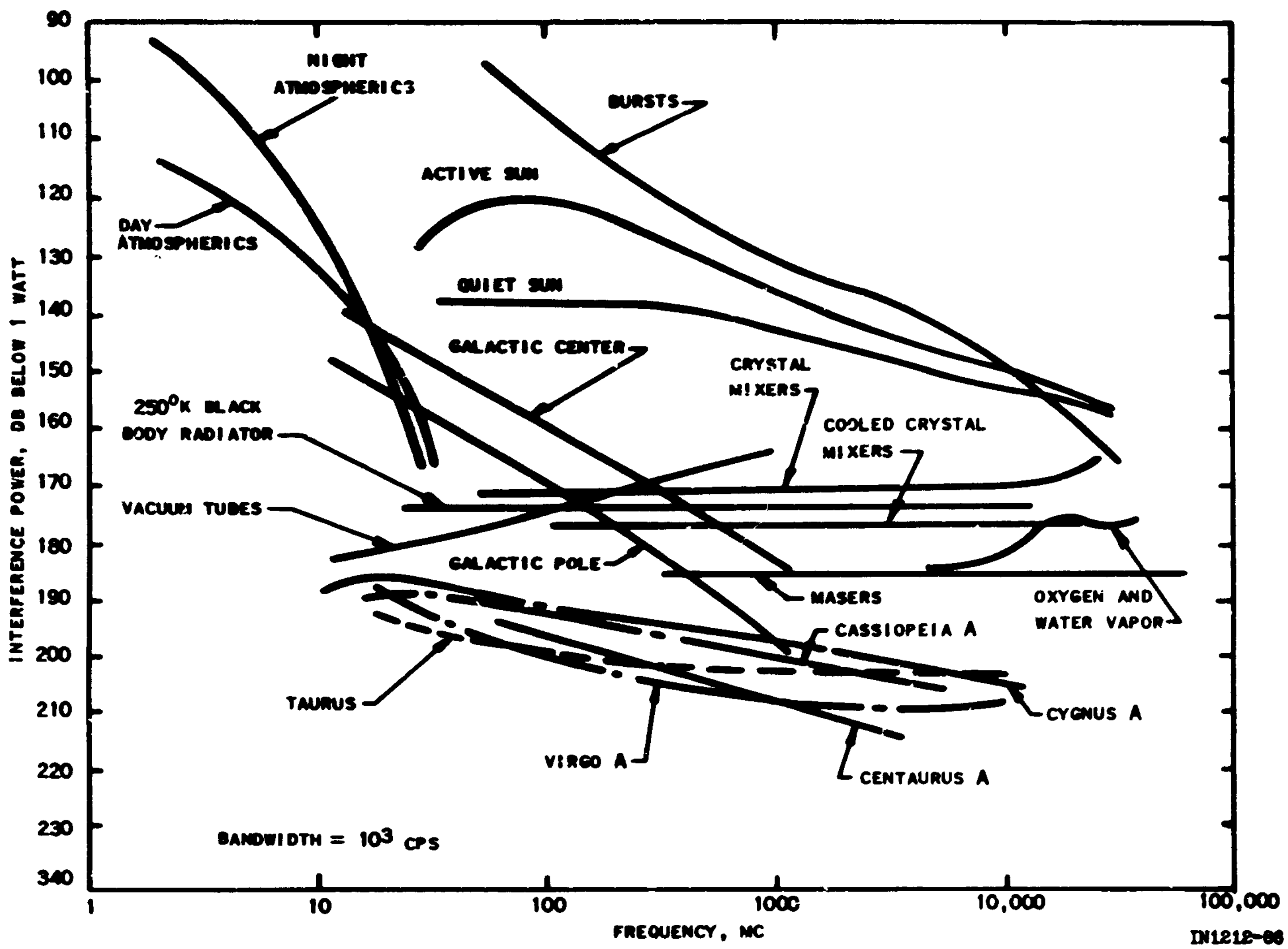


Figure 1-14. Interference Power Versus Frequency for Various Sources

b. Cosmic Interference. Cosmic interference originates beyond the earth's atmosphere and is generated, to some degree, in all areas of the universe. Like atmospheric interference it consists of short, randomly recurring pulses completely without regularity in phase or amplitude, and its energy is not confined to a particular band. The cosmic interference intensity varies with celestial longitude and frequency. On the surface of the earth, cosmic interference is effectively confined to uhf and vhf frequency regions because of absorption and reflection of other frequencies. Above the ionosphere, communications between space vehicles at frequencies below 20 mc will be subject to the full effects of cosmic interference. The true intensity range of this interference above ionosphere shielding is as yet unknown for lower frequencies. Cosmic interference is most prevalent below 250 mc, while solar interference extends beyond 30,000 mc. Major sources of galactic interference are the Milky Way and the sun. By using narrow bandwidth antennas and the highest frequency, interference from these sources can be greatly reduced. Figure 1-14 indicates typical values of cosmic interference measured by a skyward directed, high-resolution antenna.

c. Solar Interference. Solar radiation emanates from sources located in the chromosphere and corona of the sun. The sun behaves, however, very much like a black-body radiator at 6000°K for frequencies in the infrared and visible light spectrum, although it emits considerably more radiation than can be accounted for by black-body analysis at lower frequencies. The lower frequencies are generated in the corona and the higher frequencies in the chromosphere, with both contributing radiation at intermediate frequencies. Because the source location varies in surface depth, the apparent temperature for each frequency also varies accordingly. The intensity of solar radiation fluctuates with a periodicity of weeks or months and corresponds to sunspot activity. It increases during periods of sunspot activity and decreases when there is less sunspot activity. These periods of increased radiation last several days at a time. At such times, interference levels may be 10 to 20 db greater than the normal level of a quiet sun. At lower frequencies, solar radiation does not fluctuate as much as it does at high frequencies. Radiation is first noticed at the higher frequencies; afterwards, at the lower frequencies. The radiation generally occurs in bursts that are greater in intensity than the general background interference level. The radiation accompanying these storms is strongly circular in polarization. Accompanying solar flares are outbursts of interference that last several minutes and increase the interference level of the sun by many orders of magnitude. Sometimes the frequency spectrum of these outbursts is quite wide and involves the major portion of radio frequencies now in use. Figure 1-74 shows appropriate interference levels for an active, quiet, and violent sun. These levels were determined for an antenna oriented toward the sun with a beamwidth nearly equal to the solar diameter. Five different types of solar radiation are described in table 1-1.

TABLE 1-1. CLASSIFICATION OF SOLAR WAVES

| Class Characteristic | Basic Component (Thermal) | Slowly Varying Component | Short Term Interference | Unpolarized Burst | |
|---|---|---|---|---|---|
| | | | | Outburst | Isolated Burst |
| Wavelength range | Unlimited | 3-60 cm | 1-15 m | 1 cm-15 m | 50 cm-50 m |
| Duration | | Weeks or months | Hours or days | Minutes | Seconds |
| Polarization | Random | Trace of circular | Strongly circular | Random | Random |
| Place of origin | Whole sun | Number of small areas | Small area above sunspot | Small area rapid movement | Small area |
| Associated optical feature | | Sunspots and others | Large sunspots | Flare | Unknown |
| Remarks | Constant over years | 27-day component | With or without numerous bursts | No certain distinction | No certain distinction |

d. Galactic Point Sources of Interference. Discrete sources of radio interference of extraterrestrial origin have been observed since 1946. When a source is sufficiently small, it is considered a point source. Most point sources are termed radio stars, but have not yet been shown to correspond to actual optically visible stars. Usually, if a source subtends an angle less than one degree, it is considered localized, and if more than one degree, an extended source.

(1) Many radio stars have been discovered in the last decade since the advent of radio astronomy. The use of low-noise receivers increases the possibility of discovering more radio stars. Only a few of these sources, however, emit electromagnetic energy of sufficient magnitude to warrant consideration as important sources of background interference to present receiving systems. The interference detected from five of the more important radio stars is shown on figure 1-14. The values used were converted from flux density to power, assuming an effective capture of one square meter.

(2) Interference may also be contributed to background level by the mechanism of atmospheric absorption and reradiation. Two constituents of the atmosphere, water vapor and oxygen, absorb microwave energy and collectively emit absorbed energy as black-body radiation. In effect, the receiving antenna is surrounded by an absorbing medium that acts as a thermal source of radiation. The amount of such interference that appears at the antenna is a function of path length, absorption coefficient, and atmospheric temperature. The absorption coefficient also varies with atmospheric temperature, pressure, water-vapor content, and path direction. Most absorption occurs in the troposphere, where density of atmospheric elements is still significant. The interference source is therefore confined to regions of the atmosphere where moderate temperatures (approximately 300°K or less) prevail.

e. Radiation From the Earth. The apparent temperature of the earth varies with latitude, topography, surface-atmospheric characteristics, and time. Although the earth is in apparent thermal equilibrium, sections of its surface are not. The polar regions emit more thermal energy than they receive, while the equatorial regions absorb more energy than they emit. An earth satellite, maintaining communication with a ground station, experiences background interference due to the thermal radiation emitted by the earth. This interference, depending upon the frequency employed, may limit the minimum ground station power level that can be used for communication with a satellite. Figure 1-14 shows the interference power for a black body temperature of 250°K, detected by a receiving antenna with a half-power beam area equal to the total radiation surface.

1-23. Man-Made Sources of Interference

Man-made interference impedes the reliable and efficient utilization of present-day electrical and electronic equipment. To control man-made interference, it is necessary to recognize its many possible sources, the methods by which it is propagated, and its effects upon susceptible equipment. This task becomes very difficult when several different types of interference occur simultaneously. Some of the more common sources of man-made interference are described here and listed on table 1-2.

a. Electrical Machinery. Electrical machinery constitutes a serious source of broadband as well as narrowband interference. Broadband noise is generated during the commutation process by the brushes and the armature, arcing in bearings, friction between moving parts, internal arcing, and control windings. Narrowband interference arises from poor machine symmetry causing the generation of harmonic frequencies. Section X of chapter 3 discusses in detail the control of interference in electrical machinery.

TABLE 1-2. TYPICAL MAN-MADE INTERFERENCE SOURCES

| Broadband | | | Narrowband | |
|---|---|---|---|---|
| Transient | Intermittent | Continuous | Intermittent | Continuous |
| Function switches | Electronic computers | Commutation noise | Doppler-shift radar while scanning | |
| Motor starters | | Electric typewriters | Radio transmitters | Power-line hum |
| Thermostats | | | | Receiver local oscillators |
| Timer units | Motor speed controls | Ignition systems | | |
| | Poor or loose ground connections | Arc and vapor lamps | | |
| | Welding equipment | Pulse generators | | |
| | | Radar modulators | | |
| | | Sliding contacts | | |
| | | Teletypewriter equipment | | |
| | | Voltage regulators | | |

b. Frictional Static. Frictional static is the accumulation and discharge of electromagnetic energy resulting from a rubbing motion between two materials. There are different theories as to the cause of frictional static. A few of the most common are:

1) Movement of electrostatic charges which build up an electromagnetic field

2) Instantaneous change of field intensity (that is, a step function change) causing the collapse of the electrostatic field

3) A combination of the first two causes

(1) Whenever an electric field is set up in a substance, a displacement of electric charge takes place. The nature of the displacement is dependent upon the nature of the substance. The positive charge within the substance is displaced or oriented in the direction of the field intensity, and the negative charge in the opposite direction, until an opposing force is set up which just balances the forces due to the impressed field. In metallic conductors, flow of electrons in a direction opposite to the field constitutes electric current. In electrolytes, there is a migration of positive ions in the direction of the field. In insulators, the migration is negligible, but molecules of the substance are deformed and reoriented in such a manner as to produce a momentary motion of positive electricity in the direction of the field, and of negative electricity opposite to the direction of the field.

(2) Frictional static can be a serious source of interference. Frictional static may occur between metals, between a metal and a nonmetal, or between nonmetals. The degree of seriousness is determined by the amount of friction encountered. Frictional static may occur between any two moving surfaces or between the contact of two charged surfaces. Examples of this phenomena are:

1) Belt static -- a static charge between a dielectric (nonconducting) type belt and its pulleys

2) Bearing static -- occurring between bearings and their lubricant or housing

3) Tire and track static -- appearing between the tires or tracks of moving vehicles or tanks and the road

4) Gear static -- static resulting from two gears moving against each other

(3) Static charges on belts are a common occurrence in industry. These charges develop on both power-transmission and conveyor belts. Low temperatures appear to be more favorable for accumulation of these charges, although they may become serious in dry atmosphere at any temperature. Static arising from vehicular tires is quite evident in mobile transmitters and is very pronounced when a vehicle is traveling on paved roads. When two gears of similar metal mesh, little static is generated; when two gears of dissimilar metals mesh, interference in the form of pulses at the beginning of motion appear. The electrolytic action between two different metals causes an electrical discharge as the gears mesh.

c. Switches. All switches cause essentially the same types of interference whether mechanical, thermostatic, electromechanical or electronic in operation. The switch is essentially a device that can abruptly change its electrical impedance from zero to infinity or from infinity to zero. This sudden change causes rapid current and voltage transients throughout the circuit, which result in generation of broadband interference. In the case of manually actuated mechanical switches, the interference produced is generally of relatively short duration unless there are capacitors or inductors in the circuit. Usually such interference is not repetitive, or, if it is, repetition is at a slow or irregular rate. Switching interference is more severe when large current values are involved because arcing across switch contacts when making or breaking circuits greatly intensifies the interference in both level and duration.

Electromechanically actuated switches such as relays, vibrators, and buzzers create exactly the same type of broadband interference as do manually actuated switches, but usually at a faster rate. Repetitive interrupters, such as vibrators, produce the same broad spectrum interference on a continuous basis.

d. Discharge Tubes. Thyratrons and similar gaseous discharge tubes are frequently utilized for switching purposes because of their extremely fast operating time. Because of this fast operating time, however, such tubes create very substantial amounts of interference, particularly when handling large currents. Thyratrons having turn-on times in the order of 1 μsec are quite common. Just as with manual switches, thyratrons being operated infrequently cause considerable broadband interference, though only for a relatively short time duration. Thyratrons, when discharged repetitiously at a high rate, such as when used for the control of motors and other equipment, can create very intense broadband interference on a continuous basis. There are numerous devices, in addition to thyratrons, which produce the same type of interference. These are gaseous voltage regulators, mercury vapor rectifiers, cold-cathode rectifiers, and fluorescent lamps.

e. Rectifiers. Rectifiers have been described as nonlinear circuit elements having the property of passing a greater current in one direction than in another. There are two broad rectifier classifications for electromagnetic interference: mechanical and electronic. Virtually all rectifiers presently in use, or anticipated for future use, are of the electronic type; mechanical rectifiers are now obsolete.

(1) Most rectifiers produce at least two types of interference: harmonic and broadband. Harmonic interference is generated as a result of the nonlinear characteristic of a rectifier. The signals produced include the applied frequency, its modulation components, harmonics, and intermodulation products. These often extend over hundreds of megacycles throughout

the spectrum. When an applied signal is nonsinusoidal in waveform and/or of high frequency, the cutoff time of the rectifier becomes very short, and it generates broadband interference.

(2) Mercury vapor and gaseous-type rectifiers generate interference because of the manner in which they function. The continual establishment of new conduction paths through mercury within the tube and the continual breakdown of existing paths, all with considerable arcing, create interference in additon to the usual harmonic and switching type interference produced. The mercury vapor tube presents a very serious interference problem and is not recommended for general use in new equipment. When used, special precautions should be taken in placement, shielding, bypassing, and filtering of interconnecting wires.

(3) Silicon-controlled rectifiers are related to the transistor in much the same manner as the thyratron is related to the triode. These devices are capable of handling substantial amounts of power and are either nonconductive or conductive, depending upon their bias. When triggered, the controlled rectifier reaches its maximum state of conductivity very rapidly. The result is the production of a broad band of electromagnetic interference. Because controlled rectifiers are often used as repetitious switching devices in static inverters and power converters, they can be a continuous source of broadband interference. They generally, however, produce less interference than thyratrons or comparable mechanical interrupting devices. Control of this interference is best accomplished during initial design through employment of suitable shielding, filtering, and bypassing techniques.

(4) Zener diodes, because of their inherent nonlinearity, produce the same type of harmonic interference as other rectifiers.

In minimizing this interference, measures similar to those used with semiconductor rectifiers can be employed. Zener diodes generally produce more broadband switching interference then other diodes because of their unique self-bias characteristic. It is this characteristic which causes the Zener diode to switch rapidly from the nonconductive to the conductive state when a minimum voltage is exceeded, and to switch back to the nonconducting state when the applied voltage falls below the required level. Because Zener diodes are frequently employed in regulator applications where constant switching operations occur, and in ac circuits where continuous switching occurs at the power frequency, their presence indicates the need for appropriate interference suppression.

(5) Tunnel diodes are relatively new semiconductor devices which are finding extensive experimental and commercial application throughout the electronics industry. The tunnel diode can be an interference source if not properly utilized. If not used properly they have a very definite tendency to oscillate at several unrelated frequencies simultaneously. The resulting spurious radiations may or may not have any relationship to the intended frequency of operation and can create very serious interference problems. Even the use of crystal control will not assure single frequency operation because it is possible for one output signal to be at the crystal frequency, while other signals are at frequencies determined by one or more resonant peaks of the tuned circuit. The tunnel diode is also extremely susceptible to shock excitation from strong external electromagnetic fields. The design engineer must exercise care in selecting and establishing circuits for tunnel diodes. Interference control measures, such as short point-to-point wiring, adequate shielding, filtering and bypassing should be used to minimize spurious-signal generation.

(6) Varactors, though not rectifiers, are similar to semiconductor diodes in construction. When connected in circuits so as to be negatively biased and nonconducting, varactors present a capacitance that varies with applied potential. These devices serve many functions, such as variable capacitors for remote tuning of radio sets. Power varactors have also been developed for use as uhf frequency multipliers. Because they are nonlinear in operation -- and are excellent harmonic sources, as indicated by their efficiency as passive frequency multipliers -- they are potential sources of electromagnetic interference. Th types of interference that they produce vary considerably with the circuits in which they are utilized.

f. Regulators. The manner and rate at which voltage is controlled by a regulator determines the spectrum and intensity of the resulting interference. Variations in current, which result from every adjustment in voltage, cause interference to be generated over large portions of the electromagnetic spectrum. Large and rapid incremental voltage adjustments cause broader and more intense interference than slower and more gradual adjustments. This interference may be widely disseminated by direct radiation and conduction. Thus, for the design engineer concerned with interference control, there is a conflict between requirements for rapid, wide-range voltage regulation and minimal interference generation.

(1) Various types of voltage regulators are available. All have in common the inherent ability to generate interference through the very process of regulating voltage. Many regulators produce additional interference, which results from their individual characteristics. Examples include switching and arcing transients produced by reed-type regulators; harmonic and switching type transients produced by gaseous regulators, Zener diodes and other semiconductors; and interference that results from current pulses peculiar to magnetic-amplifier type regulators.

(2) Current regulators operate in the same manner as voltage regulators; they, however, sense and regulate current through a circuit rather than voltage applied to it. The basic operation of current regulators is similar to that of voltage regulators. Similarly, current-regulator interference results from individual regulation transients and from the harmonic generation capabilities of its nonlinear components. The chief difference between voltage and current-regulator interference is the magnitude: current regulators frequently involve variations in larger currents than do voltage regulators. They generate, therefore, more intense interference than do voltage regulators.

q. Corona. Corona occurs when a dielectric material is subjected to high voltage and ionizes. In high-voltage equipment, corona is most likely found in the air gap adjacent to high-voltage buses, bushings, and locations where small air gaps are present, and where the electrostatic field is nonuniform. For uniform fields, corona usually precedes arc breakdown. The corona produces broadband interference at random frequencies and is very difficult to suppress. The best method that can be used to eliminate this type of interference is to eliminate the source of corona; this can be accomplished by increasing the spacing between breakdown surfaces, improving the geometric configuration, providing shielding between breakdown surfaces, improving voltage distribution, and eliminating air gaps, either electrically or mechanically. Each of these methods is described here:

(1) Because corona in air is an ionization of air caused by a voltage high enough to exceed the critical gradient of the air gap, the first step in eliminating corona is to increase this air gap. The air gap size for a required corona level depends on geometric configuration. The curves on figures 1-15 through 1-18 show the effects of spacing of four basic configurations.

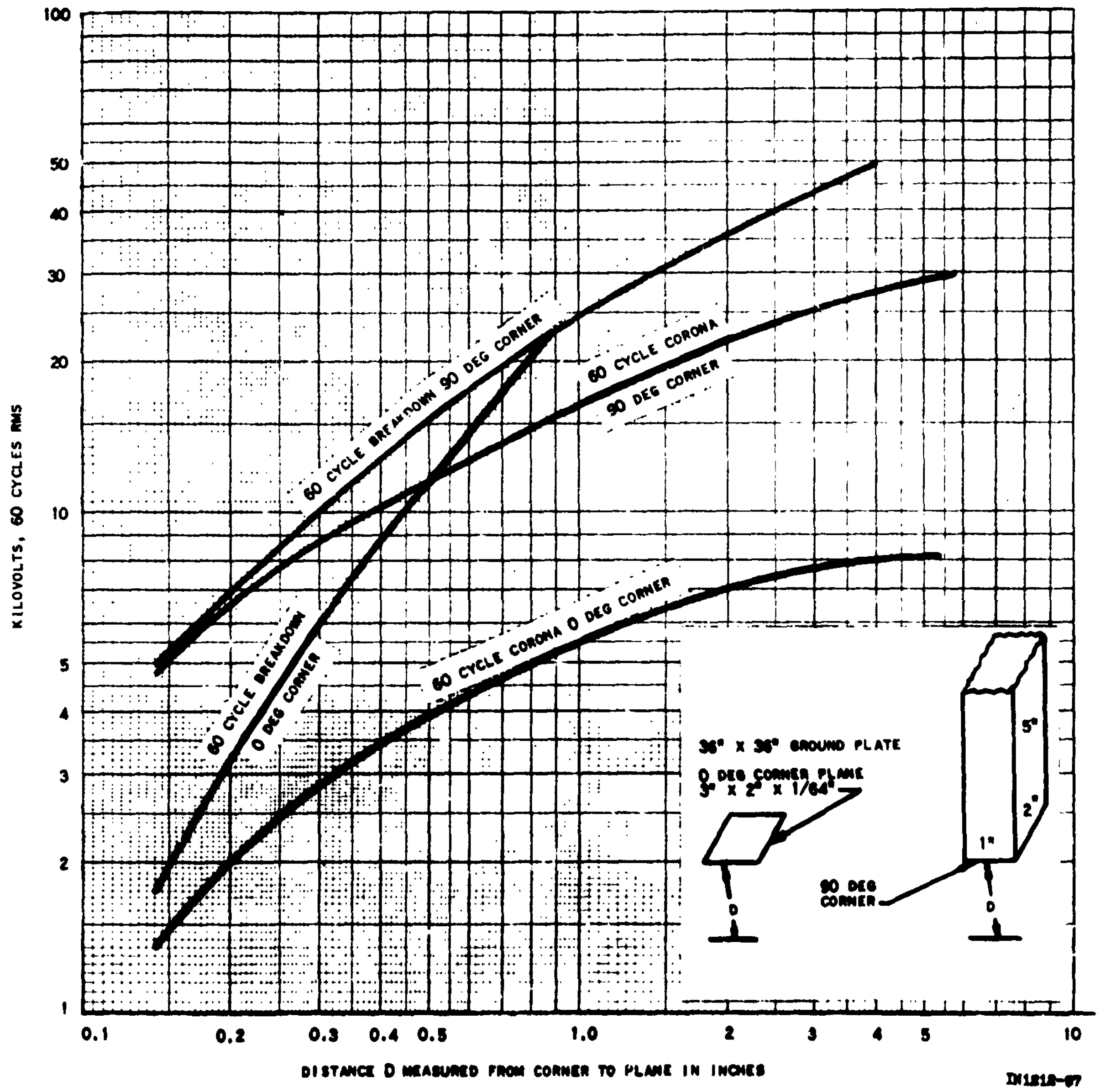


Figure 1-15. Corona Level, Plane-to-Corner Configuration

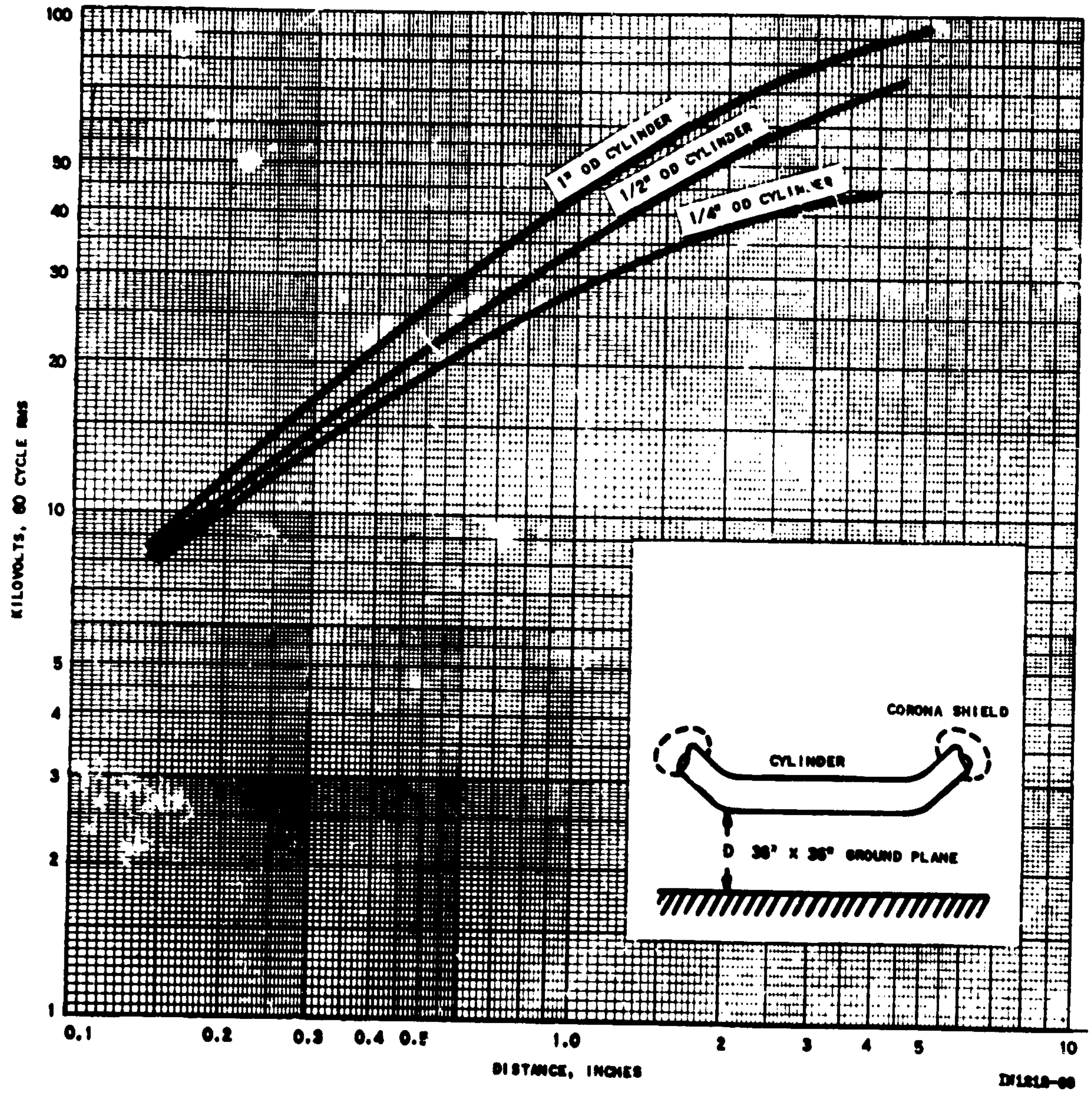


Figure 1-16. Corona Level, Plane-to-Cylinder Configuration

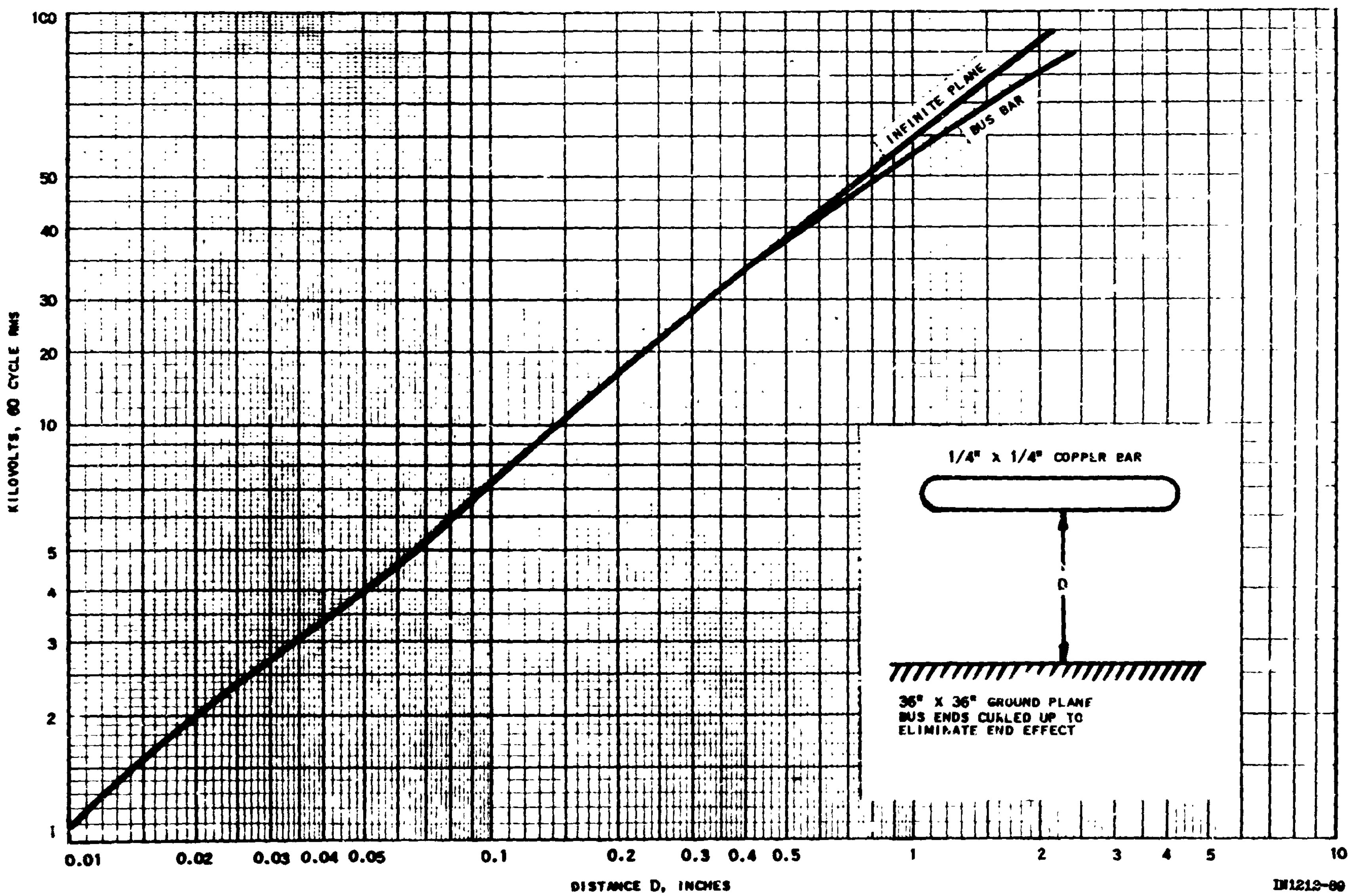


Figure 1-17. Plane-to-Plane Relationship

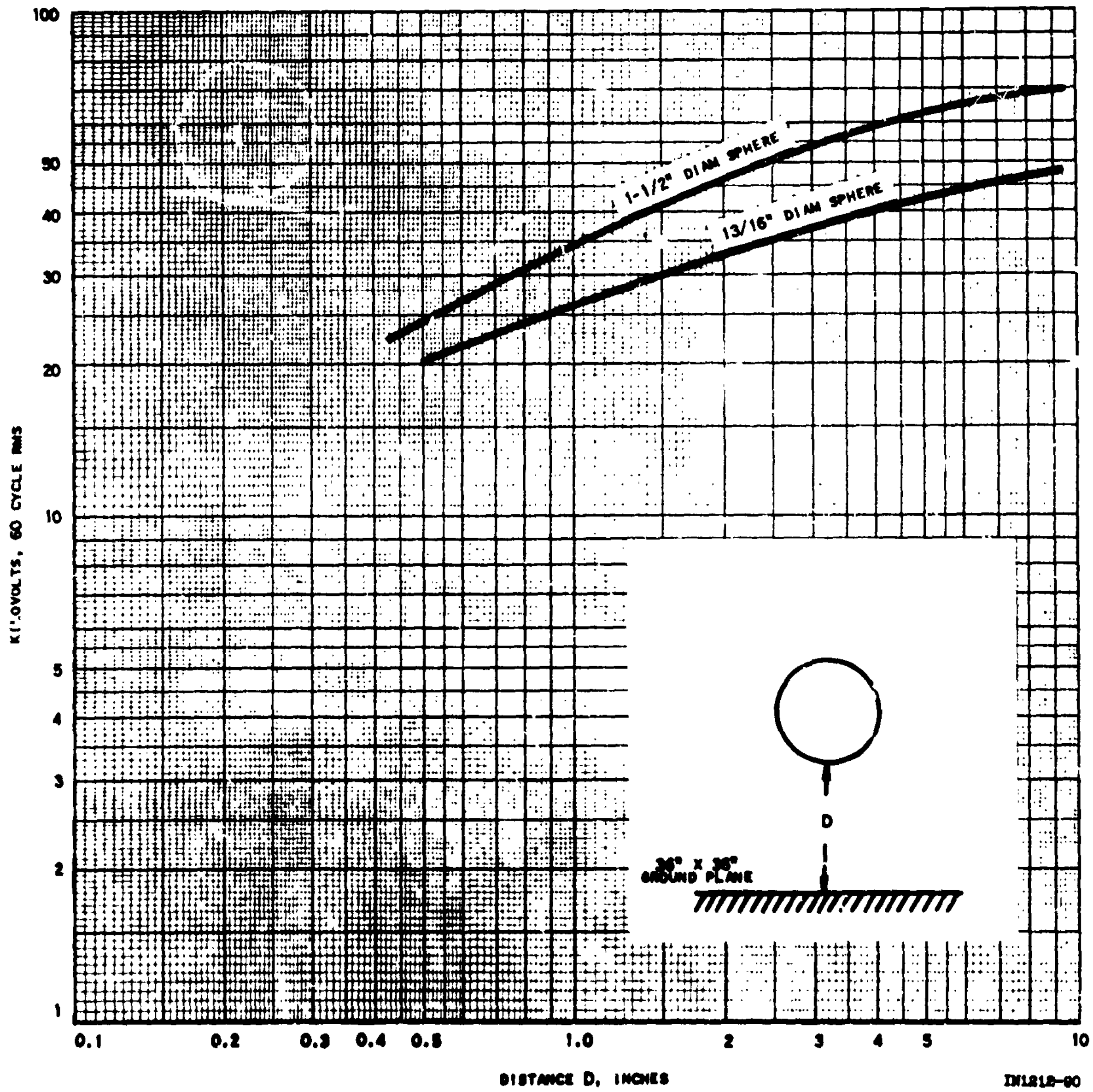


Figure 1-18. Plane-to-Sphere Relationship

(2) When space requirements make it impossible to increase the size of the air gap, other methods of decreasing voltage stress must be developed. One method, a change in geometric configuration of the device, has an important effect on changes of geometry in a one-inch air gap. The data in table 1-3 shows the effectiveness of changing the geometric configuration. In many cases, minor configuration revisions will result in a major improvement in corona level. It is good design practice to round off sharp corners on bare conductors in the vicinity of an air gap that might undergo high voltage stress.

TABLE 1-3. RELATIONSHIP OF GEOMETRY OF DEVICE TO CORONA VOLTAGE

| Geometry | Kv Corona Voltage (with 1-inch air gap) |
|---|---|
| 0 degree corner | 5.4 |
| 90 degree corner | 17 |
| 13/16-inch diameter sphere | 26 |
| 1-1/2-inches diameter sphere | 35 |
| 1/2-inch diameter cylinder | 35 |
| 1-inch diameter cylinder | 42 |
| Plane | 51 |

(3) The application of high-dielectric insulating shields around conductors that have high stress points eliminates corona. Because it is usually cheaper to resort to geometry correction, shielding is avoided where possible. However, there are some economical applications of a shield on high-voltage equipment. The shield is placed over the hardware to eliminate high-stress points at threads and corners.

(4) Usually, complicated configurations can be reduced to an equivalent circuit. The equivalent circuit of an insulated conductor could be a simple series capacitance circuit. Because the

dielectric constant is known for the insulation being used, it is easy to calculate the impedance of each capacitor. The voltage distribution across each capacitor can be calculated as a series circuit. The dielectric constant of the insulation in series with an air gap determines the voltage distribution. For a given air gap in series with a solid insulation, it is preferable to choose an insulation with a low dielectric constant so that stress on the air will be minimized.

(5) Because corona occurs when air is overstressed, it is possible to eliminate it by eliminating the air gap, either electrically or mechanically. To eliminate an air gap electrically, both surfaces bordering the air gap should be coated with a conducting or metalized paint or metal foil. The two conducting surfaces are then tied together electrically to eliminate voltage across the air gap. The air gap can be eliminated mechanically with a filler, such as transformer oil, epoxy casting resin, or petrolatum.

## Section III. INTERFERENCE GENERATION

### 1-24. General

a. Basically, all sources of interference can be divided into three groups: natural, inherent, and man-made. Natural interference is caused by atmospheric disturbances and precipitation static; inherent interference is generated within a piece of equipment (for example, thermal agitation, shot effect, and switching contacts); and man-made interference includes all other interference from external sources. Any piece of electrical or electronic equipment must be regarded as a potential source of man-made interference. The electrical or electronic device that produces an abrupt change in current, such as an electric spark or an arc discharge, may cause man-made interference over a wide band of frequencies.

b. Table 1-4 and the left-hand block on figure 1-19 indicate the typical sources of interference within an electronic complex. The central block on figure 1-19 comprises parameters and variables between the transmitting source and the receiver. It is analogous to the transfer impedance or admittance between a driving point and receiving point in electrical circuits -- that is, to the transfer medium or transmission region. This region will generally have the effect of attenuating the energy from the source before it impinges upon the receiver. The possibility of signal reinforcement from fortuitously located intervening objects in the propagation path is remote. The right-hand block on figure 1-19 refers to the receiving element, which is the element responsive or susceptible to the impinging interference. Susceptible elements are not limited to receivers alone. Any equipment sensitive to extraneous electric or magnetic fields is classed as a responsive element, including computers, sensitive relays, explosives, control devices, and receivers.

c. Interference signals are associated with time-varying electrical or magnetic fields. They are generated by the fields that result when the time rate of change of current is not equal to zero. Mathematically:

$$a)\ \frac{di}{dt} = 0 \quad \left(\begin{array}{c}\text{No}\\ \text{Interference}\end{array}\right) \qquad b)\ \frac{di}{dt} \neq 0 \quad \left(\begin{array}{l}\text{level of interference}\\ \text{present is related to}\\ \text{value of } di/dt\end{array}\right) \qquad (1\text{-}1)$$

Equation 1-1(a) is an absolute criterion for the absence of interference. Variation will cause current fluctuations that result in the generation of interference. Variations in current result from changes in voltage, impedance, a combination of both, or equivalent changes in current-operated circuits.

TABLE 1-4. CLASSIFICATION OF INTERFERENCE

| Interference Characteristics | Category | Cause or Source |
|---|---|---|
| Spectral Distribution | Narrowband | Spurious radiation, parasitic, harmonic, and other oscillations at distinct cw frequency |
| | Broadband | Periodic or aperiodic rf and voltage pulses |
| Mode of Transmission | Radiated | Any undesirable rf signals, such as unshielded ignition systems, poorly shielded radar modulators, escaping rf energy from waveguide flange couplings, commutators, relays, co-channel and adjacent-channel transmitter outputs |
| | Conducted | Any undesired signal that reaches a receiver by direct, inductive, or capacitive coupling through the antenna lead-in cable, power leads, signal leads, or control cables |
| Origin of Generation | Natural | Cosmic and atmospheric electrical disturbances |
| | Inherent | Background noise originating in the antenna, circuitry, electron tubes, and crystals, such as thermal agitation and shot effect |
| | Man-made | All interference from external sources other than natural or inherent |

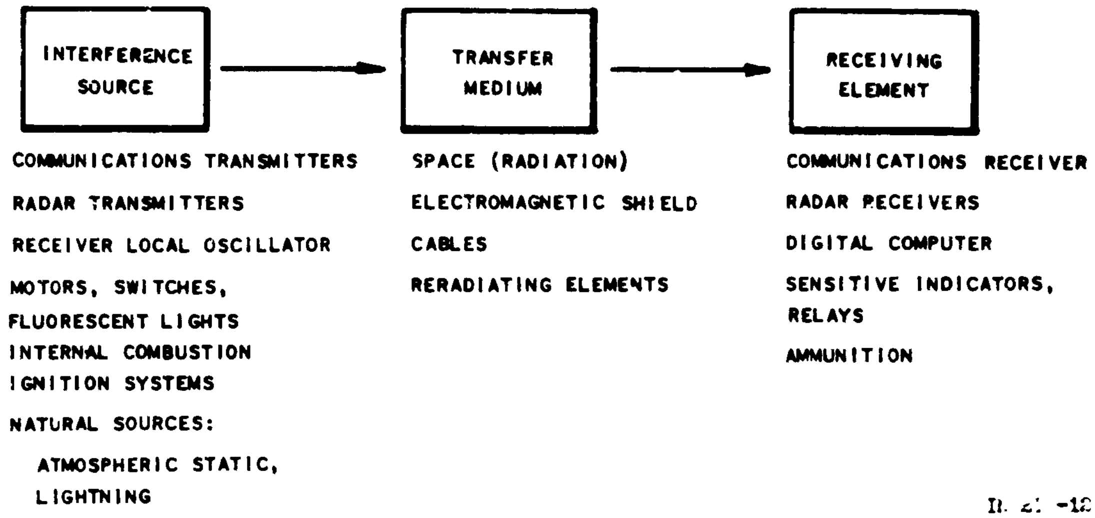


Figure 1-19. Three Basic Components of an Interference-Susceptibility Situation

## 1-25. Voltage Variations

Changes in potential or voltage arise from processes used in the generation and control of electricity. Causes of interference due to voltage variations include the relative mechanical motion between conductors and magnetic fields, switching actions, and the generation of equivalent potentials in nonlinear impedances.

## 1-26. Impedance Variation

When impedances vary in a circuit, the result is mathematically equivalent to having an equivalent variable voltage or a variable current generator in series with an internal impedance. Any nonlinear impedance is therefore a potential source of interference. When switching occurs, the magnitude of the impedance across the switch changes abruptly from megohms to milliohms. This action causes a rapid, high-level rate of change of current. These sudden changes in impedance result in the generation of much greater levels of broadband spurious energy than slow or gradual changes in impedance.

1-27. Mechanical Motion

a. Voltage is induced between the ends of a conductor when it passes through a magnetic field. The value of this induced voltage is

$$E = B \ell v \qquad (1\text{-}2)$$

where: B is the magnetic flux density

$\ell$ is the effective length of the conductor

v is the component of relative velocity perpendicular to $\ell$ and B

This action is the basic principle in design of ac and dc machines. The current does not always occur as a sinusoidal or single frequency wave. A frequency component within the current spectrum may cause a malfunction to occur in a susceptible unit, depending upon the frequency and the level of the interference. In a similar manner, the magnetic field of currents that vary in amplitude in a wire may cut the wires of another circuit and induce interference voltages. The level of interference generated depends upon factors such as rate of change of current, power levels, and the existing shielding and physical layout. It is also possible that interference will be generated in wires existing in a magnetic field that are subjected to vibrational forces.

b. Severe interference is encountered with rotating machinery when commutators or slip rings are used. In a dc generator, for example, broadband interference emanates from the brushes as they contact the commutator and switch the load from segment to segment while current is being drawn. The resultant interference is caused by the change in impedance every time the current path of the commutator is interrupted. This change also occurs with slip rings, and the arcing can be observed.

1-28. Switching Action

Broad spectrum interference results from the operation of such devices as commutators, mechanical switches, electric ignition systems, fluorescent lights, gaseous and electromechanical voltage regulators, vacuum tube switch circuits, semiconductor switch circuits, cold-cathode rectifiers, and mercury rectifiers. All have one characteristic in common: they cause rapid modulation of an electric current when the contacts of the switch open, or the device cycles; that is, goes from the conductive to the nonconductive state and back again.

## Section IV. INTERFERENCE TRANSFER MEDIA

### 1-29. General

Interference may be transferred from one point to another either by conduction or radiation. Conducted interference is transferred over common interconnections between a source and a receiver, and its propagation conforms to conventional circuit theory. Radiated interference is transferred via an electromagnetic field produced by a signal source, and its distribution may be predicted by field theory (fig. 1-20).

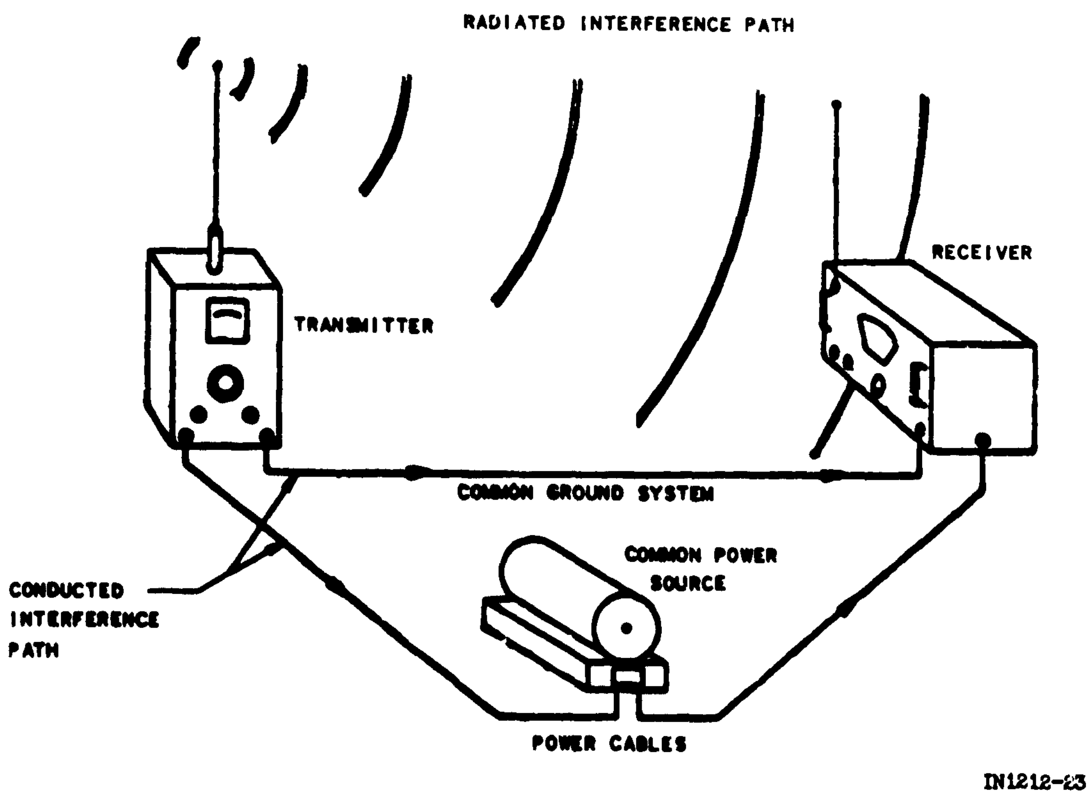


Figure 1-20. Propagation of Conducted and Radiated Interference

### 1-30. Conducted Interference

a. Conducted interference is the interference that propagates through a metallic conductor such as wiring or any metallic structure. The conductor may be a prime source of conducted interference if switching occurs in series with it, or it may be the transfer medium through which interferency may propagate.

b. Conducted interference requires a complete circuit path between an interference source and a receiver, such as may be formed by wiring, chassis, or structure. A common path for conducted interference is a ground plane. Such interference is especially severe when several electronic devices are connected to a common power source that has insufficient rf isolation. Such a set-up can result in interference being transferred to input terminals of each device because of mutual impedance in the ground path.

c. Whenever there is a direct connection between two circuits, and in addition a return path, a conduction current may flow between the circuits. The return path may be another metallic lead, a mutual capacitance, or a common ground return. The magnitude of resulting current depends on the difference of potential to ground between the points of exit and entry in the exciting circuit and on the total loop impedance between these two points. An example of interference coupling by conduction is shown on figure 1-21 which illustrates the interference from a source (motor) being transferred to a receiver through wires that are connected to a common power supply.

1-31. **Radiated Interference**

**a. General.** **Radiated interference is propagated according to the same laws of electromagnetic wave propagation that govern the propagation of desired signals. The interference field comprises three components: the static dipole field, the induction field, and the radiation field. These components vary as $\frac{1}{d^3}$, $\frac{1}{d^2}$, and $\frac{1}{d}$, respectively, where d is the distance from the source to the point of observation. The separation of the total field into its components is only an analytical process, while the physical measurement of the interference field encompasses a measure of the total field.**

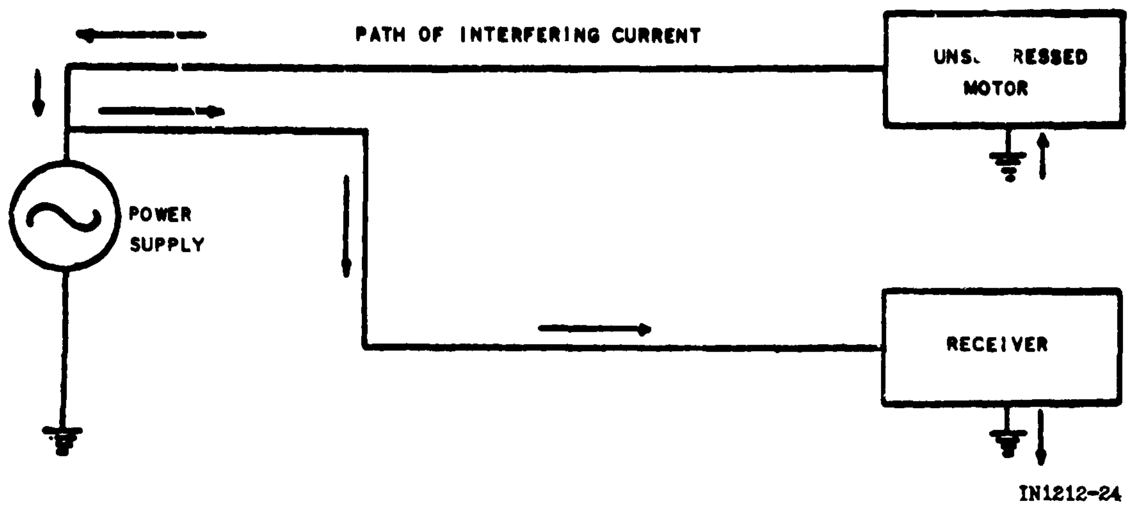


Figure 1-21. Interference Coupling by Conductor

b. Circuit Coupling. Two circuits are said to be coupled when currents or voltages in one produce currents or voltages in the other. According to circuit concepts, two circuits may be coupled either by a mutual impedance or by a mutual admittance. A mutual impedance exists between two circuits when the current flowing in circuit 1 produces a voltage in circuit 2. The magnitude of mutual impedances is the ratio of open-circuit voltage of circuit 2, with all other voltage sources removed, to the current in circuit 1. A mutual admittance exists between a point of circuit 1 and a point of circuit 2 when voltage between point 1 and some reference point (ground) produces a current to or from point 2, with this point connected to the same reference point. The magnitude of the mutual admittance is the ratio of the resulting current at point 2 to the voltage at point 1.

(1) Unless there is perfect shielding, there will be a capacitive reactance between two conductors in close proximity. The effective capacitance will vary with the distance between the conductors, their size, and the frequency of the interference potential. As the frequency of the interference signal increases, the effect of capacitive coupling becomes more pronounced.

(2) Inductive coupling refers to the process by which interference is transferred by means of the magnetic field. An alternating current in a conductor will generate a magnetic field of the same frequency around the conductor. The alternating field will, in turn, induce a signal in any closed circuit it encompasses, in proportion to the rate of change of the field.

## Section V. ELECTROMAGNETIC COMPATABILITY CONTROL PLAN

### 1-32. General

The following Electromagnetic Compatability Control Plan is a typical example of a good rfi control plan submitted to the Department of the Army by an electronic manufacturer.

### 1-33. EMC Considerations for Radio Set AN/GRC-( )

This design plan will deal with the problem of rfi in detail, discussing the types of interference expected, its sources, steps to be followed in its elimination for conformance to Military Specification Mil-I-11748, and the degree of success anticipated. Each major potential trouble source, such as power supplies, power amplifier, antenna coupler, modem and remote control unit, is treated individually.

REPORT NO.____________________

DATE__________________________

ELECTROMAGNETIC COMPATIBILITY

CONTROL PLAN

FOR

RADIO SET AN/GRC-(   )

CONTRACT NO.__________________________

1-34. RFI Design Plan for Radio Set AN/GRC- ( )

a. General. During the AN/GRC- ( ) design phase, particular attention will be given to the problem of rfi reduction. The power supply will present a difficult rfi problem because of the square wave amplitudes and distorted wave forms which are characteristic of a high frequency solid state supply. Preliminary calculations using formulas and data from Mechanical Design for Electronics Production, John M. Carroll, McGraw-Hill 1956, and commercial brochures from manufacturers of rfi shielding materials, indicate that, with proper design, interference reduction to levels specified in MIL-I-11748B can be achieved.

b. Mechanical. In the mechanical section of the design plan, the material for the four drawers is proposed as 0.125 inch thick aluminum. This choice was made to satisfy the need for rugged construction consistent with the emphasis on weight reduction. A plot of attenuation versus frequency and chassis thickness is shown on figure 1-22. The worst case frequency present in the AN/GRC- ( ) is 20 kc. From the graph, a reduction of 140 db can be expected from the chassis. In practice, this reduction is considerably less because of the openings required, and it becomes important that all interruptions in the chassis continuity be designed for maximum rfi insertion losses.

c. Detailed Discussion.

(1) Worst case analysis. The maximum square wave amplitudes present in the AN/GRC- ( ) will be 320 volts at 4 kc in the 27 volts supply, the intermediate voltage supply, and the filament supply. The fifth and higher harmonics fall within the scope of MIL-I-11748B and represent a broadband interference source of considerable magnitude. The highest interference level is calculated to be approximately 200 db above 1 microvolt per megacycle bandwidth at 20 kc. An analysis of the radiated and conducted interference reduction is shown in figures 1-23 and 1-24.

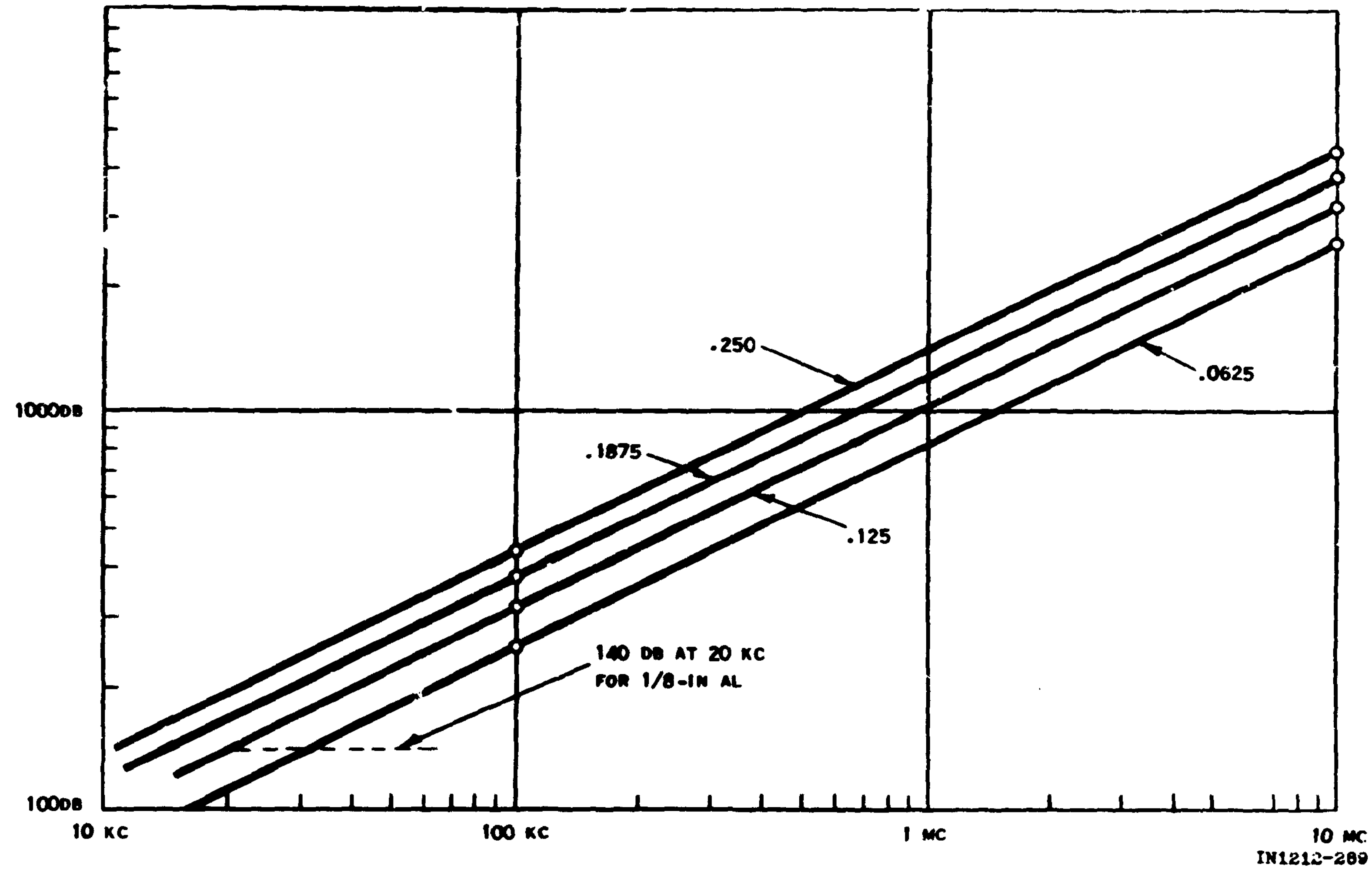


Figure 1-22. RF Attenuation vs Frequency For AL Chassis 1/16 To 1/4 IN Thick

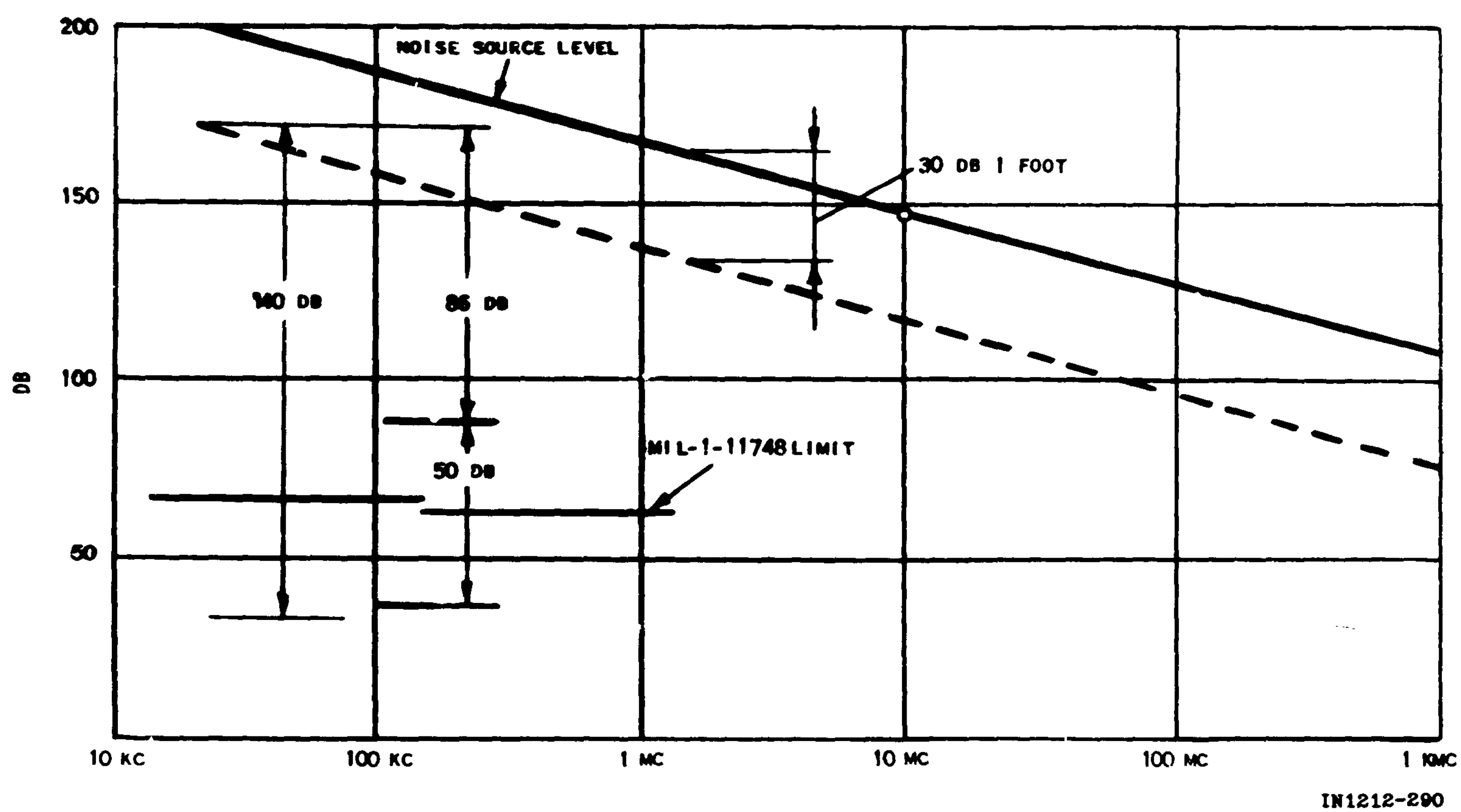


Figure 1-23. Radiated Interference Analysis -320 Volt Square Wave

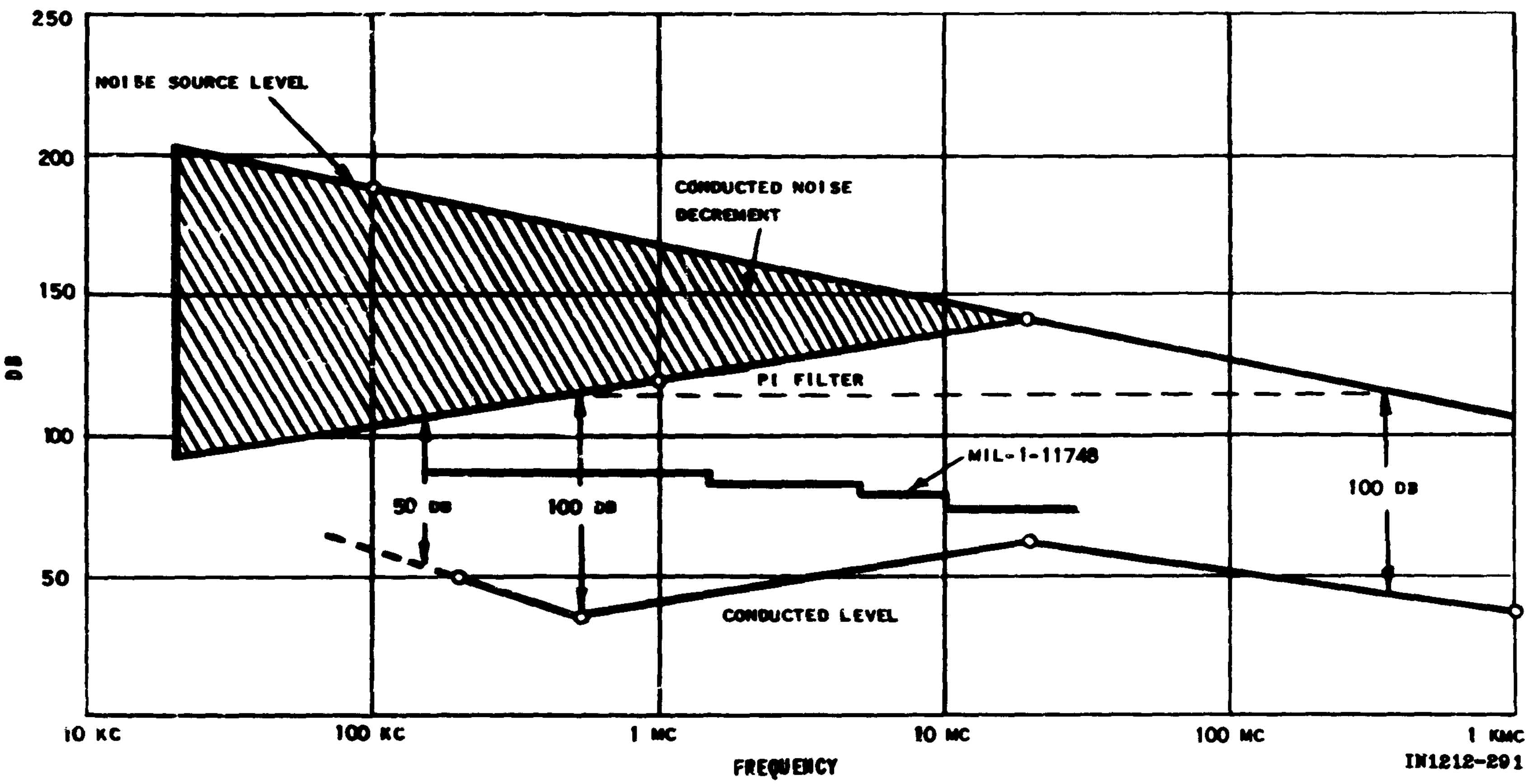


Figure 1-24. Conducted Interference Analysis -320 Volt Square Wave

Referring to figure 1-23, the first consideration is the attenuation of the chassis as the filament supply module in the power amplifier section is mounted with its open side against the section wall. The chassis reduces the interference level 140 db (from 173 db to 33 db above 1 microvolt) at 20 kc, thus meeting MIL-I-11748B specifications insofar as direct radiation through the chassis wall is concerned. However, inside the power amplifier chassis, reduction is less as the module has openings where insertion losses are considerably less than 140 db. The module is mounted to the chassis wall with a screw spacing of 1.5 inches. Figure 1-25 indicates an insertion loss of 86 db; thus the source is reduced to 87 db above 1 microvolt inside the power amplifier chassis. The chassis openings are as follows:

1) Top cover (2 inch screw spacing with mesh) insertion loss 97 db
2) Air exhaust screen (honeycomb filter with 1/8-inch cells and 1/2-inch thick) insertion loss 109 db (70 - 100 db with mounting losses)
3) Panel meter (double shielded) insertion loss estimated 100 db

From the above, it can be safely assumed that the rfi integrity of the power amplifier chassis is such that an additional reduction in excess of 50 db can be expected, reducing the interference source to 37 db, well within MIL-I-11748B specifications.

(a) The 27-volts supply and the intermediate voltage supply will be double shielded by virtue of their modularized construction; and the inverter chassis and similar rfi components will be treated so that acceptable rfi levels are obtained.

(b) An analysis of the conducted interference reduction is shown in figure 1-24. The first reduction is the conducted noise decrement (assumed losses without filtering). The unfiltered losses reduce the noise source to levels within the capability

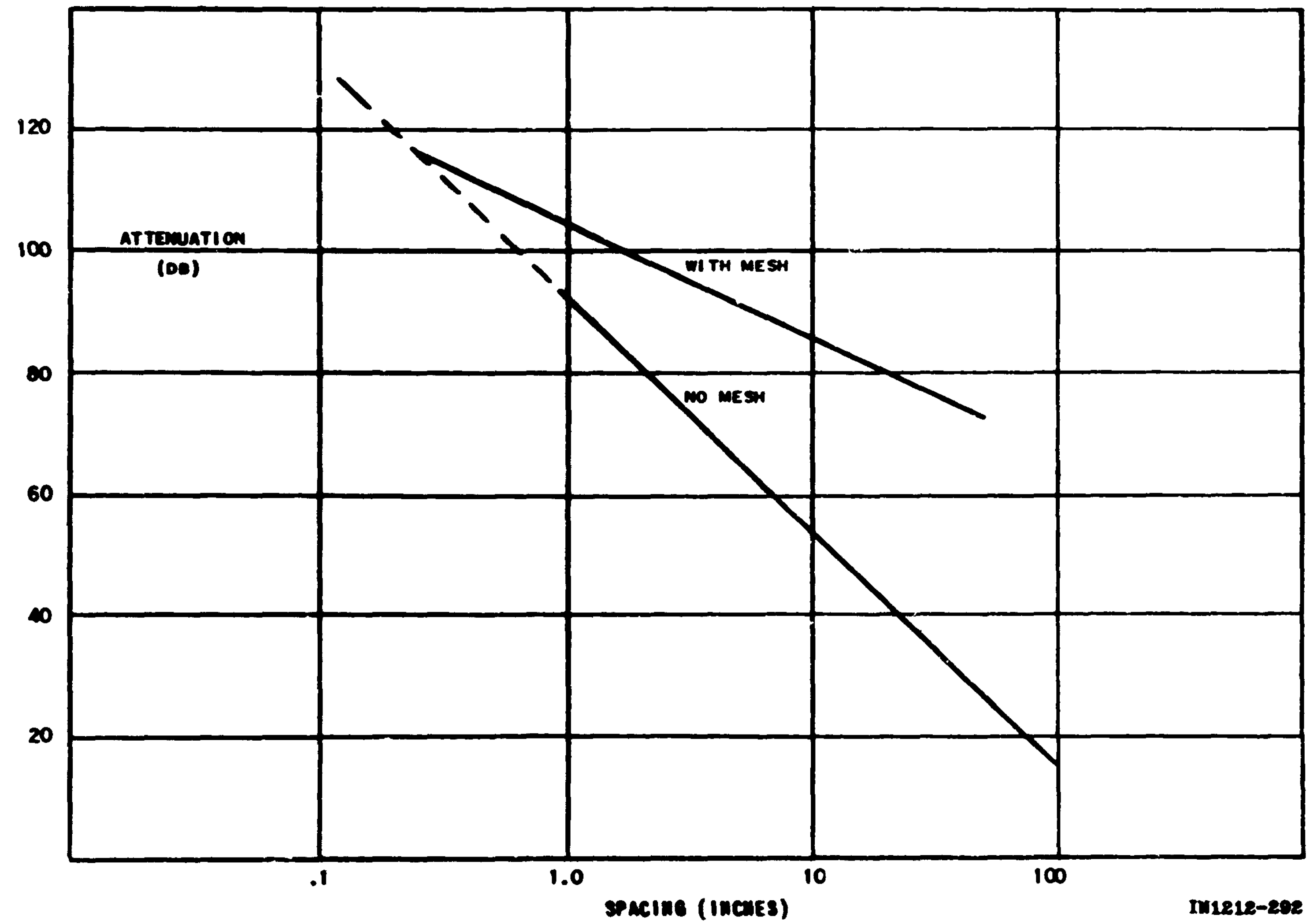


Figure 1-25. Shielding Effectiveness vs. Screw Spacing

of $\pi$ type interference filters. It is planned to use filters with an attenuation of 100 db at 150 kc (XYZ Co. $\pi$ type FSR or equivalent). This results in the conducted noise level shown in Figure 1-24 and is well below Mil-I-11748B.

(2) Insertion loss considerations - air vents. It is noted that a recommendation to use honeycomb louvered filters in lieu of screening has been made. In some cases it is not feasible to do so because of space requirements. By comparison, honeycomb filter with 1/8-inch cells and a thickness of 1/2 inch has an insertion loss of 109 db less the insertion loss of the mounting, which reduces the effective shielding to 70 - 100 db. An aluminum screen 0.060 inch thick with 3/16-inch holes and 40 percent opening has an insertion loss of approximately 85 db. The attenuation of screening 0.090 inch thick, the material which was contemplated in the design plan, should approach 100 db. Where space requirements allow, honeycomb filters will be advantageous for minimum air flow resistance and rfi losses; however, in some few areas it will be necessary to use the screening as mentioned.

d. Susceptibility. It is noted that there is no specific requirement concerning susceptibility per SCL 12345. However, the rfi precautions taken in the design of the AN/GRC-(   ) will do much to improve the susceptibility characteristics of the system.

e. RFI Reduction - Design Techniques

(1) Antenna Coupler RFI Reduction

(a) Interference sources

1) Switching transients

2) Motor drive transients

(b) Electrical design review.

1) All control leads will enter the drawer through a filter bcx with series chokes and bulkhead mounted capacitors

2) Control leads will be routed to maintain maximum distance from rf leads

3) Termination points of control leads will be bypassed to ground except where circuit functions are affected

4) RF ground leads will be as short as possible

5) Proper grounding techniques will be used

6) Relay coils will be bypassed

(c) Mechanical review.

1) The rear cover will be attached to the main chassis with machine screws spaced 1.5 inches apart (maximum)

2) The front panel will be attached to the main chassis with screws spaced four inches apart (maximum) and rfi gasketed around the periphery

3) The top cover will be attached to the chassis with 3/4-turn fasteners spaced three inches apart and rfi gasketed around the periphery

4) The manual drive shafts will enter a bearing plate located approximately 0.125 inch behind the front panel. Electrical contact from the front panel to the bearing plate via a circular section of finger stock or gasketing will enclose the shafts and prevent leakage

5) The counter windows will be made of conductive glass

6) The chassis air exhaust holes will be limited in size to 0.187 inch diameter. Honeycomb filters will be used if space permits

7) The air intake port will be screened

8) Panel meters will be shielded and meter leads filtered

(2) Power Amplifier RFI Reduction.

(a) Noise sources.

1) 320-volt, 4 kc square wave in filament supply module

2) Relay and motor drive transients

(b) Electrical design review.

1) All power and control leads will enter the chassis through a filter box with series chokes and bulkhead mounted capacitors

2) The power amplifier will be keyed by turning on the plate and screen supplies in the power supply drawer. No switching function will occur when the power amplifier is keyed on

3) Leads carrying the filament power to the tubes (12 volt square wave) will be shielded with shields grounded at both ends

4) All rf ground leads will be as short as possible.

5) All relay coils will be bypassed

6) The power amplifier plate supply lead will be bypassed near the stator

7) The pusher-driver stages will be isolated from the power amplifier stage

8) Leads carrying Average Power Control (apc) and Peak Power Control (ppc) information will be shielded

(c) Mechanical design review.

1) The filament supply module will be mounted with the open side against the power amplifier side wall with a 0.500-inch flange and 1.5-inch screw spacing

2) The discriminator/apc-ppc circuitry will be enclosed in a shielded case

3) The chassis covers will be attached with 3/4-turn fasteners spaced 3 inches apart and double core rfi gasketing will be used to ensure electrical continuity around the periphery

4) Air exhaust ports may be covered with 0.090 aluminum screen with 0.187-inch holes. The screen will be attached to the chassis walls with screws spaced 1.5 inches apart. Honeycomb filters will be used if space permits

5) Access panels will be attached to the chassis with 3/4-turn fasteners spaced three inches apart and gasketed with double core rfi gasketing

6) The access panel covering the recessed control panel will be hinged to the front panel and gasketed

7) All exposed front panel controls will be gasketed

8) Finger stock will be used to assure closure between the pusher-driver chassis and plenum

9) The panel meter will be shielded and meter leads shielded

(3) <u>Prime Power Converter Drawer RFI Reduction</u>.

(a) <u>Interference source.</u> Waveforms contained in this unit range in fundamental frequency from 47.5 to 420 cps. The silicon controlled rectifiers are used to maintain constant effective output voltage over the plus and minus 15 percent voltage variation. This implies control of the conduction angle of the SCR's. Depending on input ac voltage, the conduction angle is 170 degrees or less. However, conduction angles are kept large enough to be consistant with reasonable power factor requirements. Current obviously flows only during the conduction periods of the SCR's. The average value of the current in the worst case is around 40 amperes, depending upon what functions are taking place at that time.

(b) <u>Electrical review</u>.

1) The area of noise generation is limited to the two SCR's on the input. RFI filter of the cylindrical bulkhead mounting

type will be used. This type of filter is to be used on both input and output terminals. Suitable precautions will be taken in connecting between the connector and the filter to ensure the effectiveness of the filters

2) All other control circuits and service circuits entering or leaving the cabinet will be filtered with rfi filters having similar attenuation characteristics

(c) <u>Mechanical review</u>

1) As noted in the main design plan, efforts are being made to keep the chassis totally enclosed

2) A modification of the original mechanical design plan exists in that the prime power SCR's and diodes have moved inside the chassis entirely, rather than allow the studs to project outside the package

3) The access or service plate will have rfi gasketing material around the entire perimeter. The circuit breaker will be located within the enclosed chassis

4) A mechanical linkage will afford control over the circuit breaker from the front panel. A test panel, required by maintainability will be covered by a removable plate. The perimeter of the plate will be covered with rfi gasketing

4) <u>Main Inverter Chassis RFI Reduction</u>

a) <u>Noise sources</u>

1) High voltage inverter - sine wave inverter type power supply - converting a nominal 167 volt dc input into 2700 volts at up to 1.5 amperes. The inverter has a basic oscillation frequency of 9 kc. The power output control uses a repetition rate control system for regulation which varies the oscillation from 400 cycles to 9 kc depending upon load demands. The sine wave character of the power delivering portion of the waveform minimizes the rfi generating character of the power switching waveforms

2) Intermediate voltage supply - 320 volts, 4 kc square wave with broadband interference levels shown in figure 1-23

3) 27-volt dc supply - same as intermediate voltage supply

4) Miscellaneous circuitry - susceptibility points which are not sources, but require shielding as protection against interference from above sources

(b) Electrical design review.

1) The rectifier sections of all supplies will employ fast switching diodes to minimize the rfi content in the diode reverse recovery time

2) All leads entering the chassis will be filtered with bulkhead filters

3) Leads carrying high level current will be routed away from control leads

4) Critical control leads will be shielded

(c) Mechanical design review.

1) The interference levels produced by the high voltage inverter can be adequately reduced by the chassis

2) The intermediate voltage supply and 27-volt supply will be modularized in individual metal boxes and double shielded by virtue of the chassis

3) The top and bottom covers will be attached with 3/4 turn fasteners, spaced three inches apart and gasketed

4) The front panel meter will be shielded

5) Required panel controls will be mounted on a recessed panel covered by a hinged door which will be rfi gasketed

6) Low level control circuitry will be modular in metal boxes

7) The air intake port will be shielded with a honeycomb filter

8) Exposed front panel controls will be properly gasketed

1-35. RFI Design Plan for Modem Unit MD- ( )

a. General. The MD- ( ) Modem unit is a frequency shift keyer and converter which converts on-off information from the teletypewriters into audio tone information. The audio tones are applied to the microphone input of the RT- ( ) Receiver-Transmitter unit. It also receives audio information from the RT- ( ) and processes it to provide on-off information which is recognized by the teletypewriters as letters or characters. The unit has connectors which are the inputs and outputs of the RT- ( )'s, the speaker, microphone, handset, power, teletype inputs and outputs.

b. General Precautions. Throughout the design of the MD- ( ), rfi has been a major concern, and a number of general precautions have been taken to reduce the interference to levels which fall within the specification. These general precautions include a three-section filter box which covers the rear of the connectors. The three sections separate the power leads, the teletype loops, and the audio and control leads. Any leads which do not directly enter the main portion of the case, do not leave the filter box, but are connected between connectors within the filter box. Leads which enter or leave the main part of the case are filtered to provide adequate protection against conducted interference.

c. Individual Shielding. Radiated interference is suppressed by shielding of individual modules, and by the shielding afforded by the metal case. In the analysis that follows the shielding effect of the case and the protective measures used to shield the scope face opening are treated separately. This permits mention of the fact that the MD- ( ) is identical in construction to the RT- ( ) of the AN/GRC- ( ) system which has previously passed radiated interference specifications. Any degradation of performance must, therefore, be associated with the opening provided for the scope face. A meter shield will also be provided if it is deemed necessary.

d. Analysis. The major noise sources are listed below:

| Noise Source | Estimated Level | Remedy |
|---|---|---|
| Loop batteries (two) | Narrowband amplitude: | Shielding |
| Type: Square wave | 145 db above 1 uv at 15 kc | of case |
| Freq: 5 kc | Wideband amplitude: | |
| Amplitude: 89 volts | 191.5 db above 1 uv/mc at 15 kc | |
| Oscill. power supply | Narrowband amplitude: | Shielding |
| Type: Square wave | 163 db above 1 uv at 15 kc | of case |
| Freq: 5 kc | Wideband amplitude: | |
| Amplitude: 655 volts | 209 db above 1 uv/mc at 15 kc | |
| Mark & Space Oscillators | Narrowband amplitude: | Shielding |
| Type: Square wave | 138 db above 1 uv at 221 kc | of case |
| Freq: 221 kc & 227 kc | Wideband amplitude: | |
| Amplitude: 12 volts | 151 db above 1 uv/mc at 221 kc | |

(1) Case shielding. Analysis of shielding afforded by the case which is 0.090 aluminum includes a fixed 30 db of attenuation due to the three-foot separation between the measuring instrument and the equipment, and the losses due to penetration and reflection in the cases. Table 1-5 indicates the total losses versus frequency due to these three factors. The radiated loss takes into consideration only the case. Since the equipment is contained in covered modules, the actual interference should be lower. Data given pertains to worst case condition of voltage and load.

Table 1-5. Loss Due to Case Versus Frequency

| Frequency | Losses (db) | | | Total |
|---|---|---|---|---|
| | Fixed | Reflection | Penetration | |
| 10 kc | 30 | 59 | 28 | 118 |
| 100 kc | 30 | 188 | 90 | 308 |
| 1 mc | 30 | 595 | 286 | 911 |
| 10 mc | 30 | 1880 | 900 | 2820 |
| 100 mc | 30 | 5950 | 2860 | 8840 |

(a) Figures 1-26 and 1-27 are predicted curves of the data presented above and show the total interference level. The solid line at the lower left shows anticipated level after shielding.

(b) The scope face opening is a 2-inch diameter hole. The scope shield is a 2-inch diameter, 6-1/2-inch long pipe which, when considered as a wave guide operating below cutoff, provides 142 db of attenuation. For both broadband and narrowband interference, this is sufficient attenuation to meet the specification. Since previous calculations have been on a worst case basis, screening will be employed only if it is found to be necessary. The screening considered is steel 0.030 wire with 1/2 inch spacing which is 88 percent open and provides 24 db of shielding.

(2) Conducted interference. Figure 1-28 shows actual measurements of conducted interference level with and without filters. The measurements were made on the breadboard, and indicate additional filtering may be required at the low frequencies. Before leaving the unit, all lines pass through a filter box containing a $\pi$ section low pass filter with a 200 kc cutoff frequency. Power lines and teletype loops contain low pass filters with a 100 kc cutoff frequency. For the teletype loops, this ensures faithful reproduction of the teletype pulses.

(3) Susceptibility

(a) To self generated signals

Although the teletype sensitivity is -50 dbm, the protective measures taken predict adequate filtering for the modem. The breakdown is:

| Power Level | +60 dbm |
|---|---|
| Shelter Shielding | 20 db |
| Cable Shielding | 20 db |
| RFI Filters | 40 db (min at 2 mc) |
| TTY Channel Filters | 40 db |
| Total Attenuation | 120 db |

The input to the teletype channel is therefore, -60 dbm. The above figures again represent worst case conditions, therefore, system performance should show improvement.

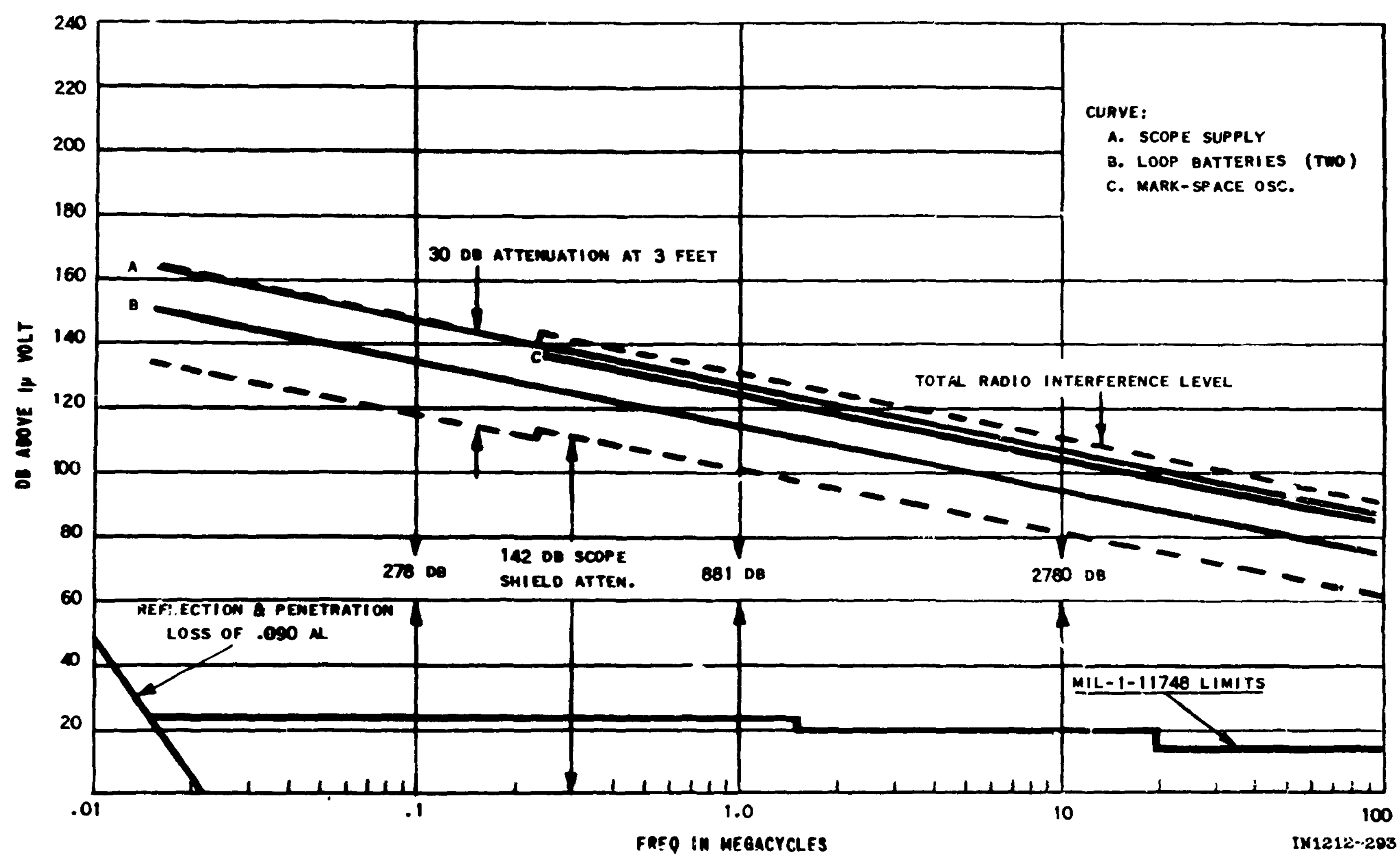


Figure 1-26. Narrow Band Radiated Interference

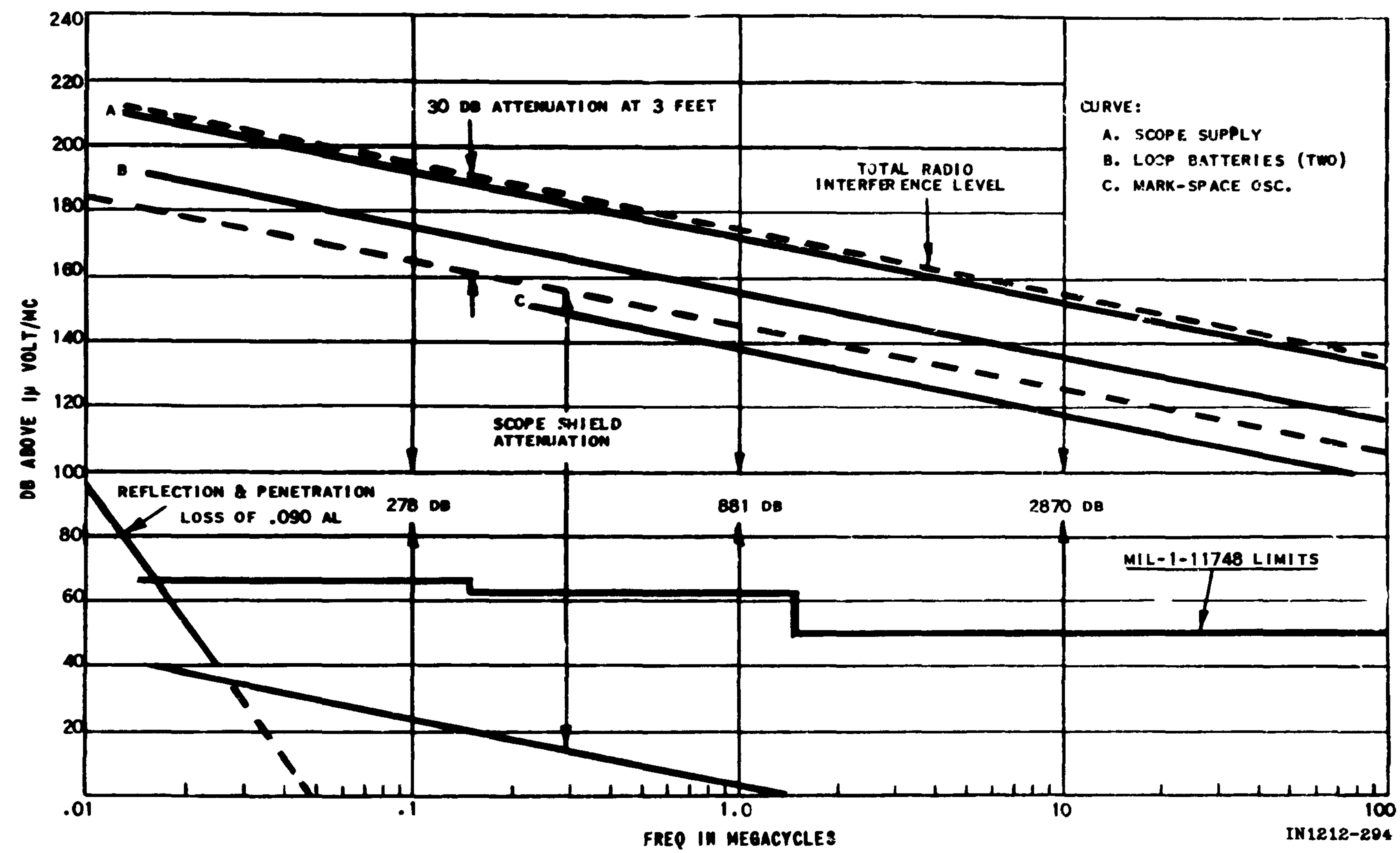


Figure 1-27. Broadband Radiated Interference

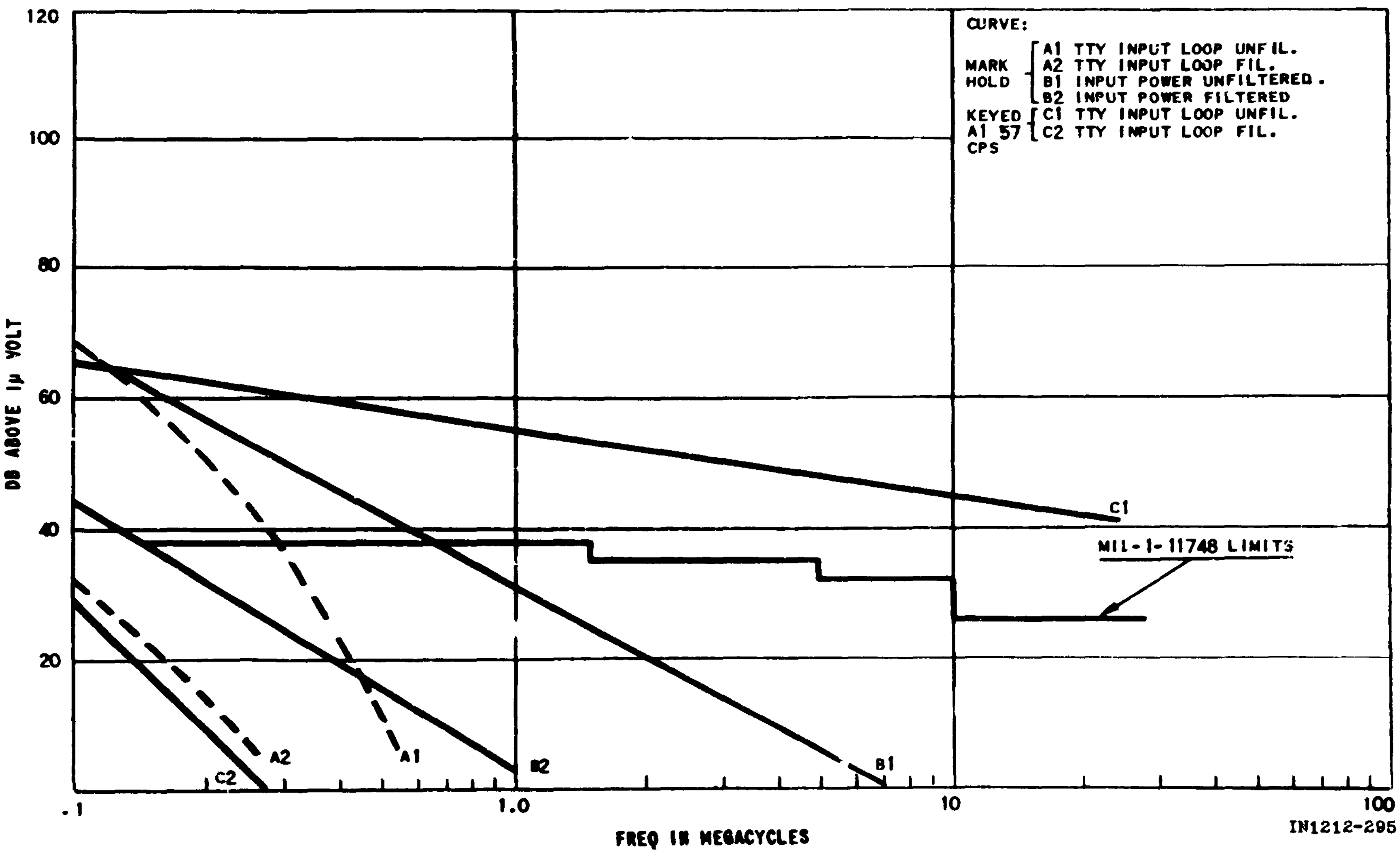


Figure 1-28. Narrow Band Conductive Interference

(b) Power line susceptibility

There will be no problem from the input power line to the modem since the supply is regulated and virtually interference free.

(c) Antenna terminal

Not Applicable

(d) Susceptibility to radiation

No undesirable response will occur with the modem under the influence of a strong electromagnetic field.

1-36. RFI Design Plan for Shelter Blower - Power Supply

a. Nature of RFI Anticipated. The rfi expected is of the broadband nature. The power is three-phase, derived from a 2400 cps ring counter. This ring counter produces a series of square waves of about 30 volts amplitude and a one microsecond rise and fall time. These pulses are power-amplified and used to drive silicon-controlled rectifiers in a parallel inverter arrangement. The resultant square wave output has an amplitude of 320 volts peak-to-peak. Thus, the rfi results from square wave voltage of 30 volts, 2400 cycles per second, and 320 volts peak-to-peak at 400 cycles per second. The 320 volts appears thrice in a three-phase sequence The 30 volts, 2400 cycle per second square wave is the lesser of the two interference generators by a factor of 10, and it is entirely within the confines of the blower supply chassis; the 320-volt square wave is fed to the blower motor in another part of the shelter. This is the significant interference of the broadband spectrum.

b. Reduction of RFI.

(1) Refer to Figure 1-29 entitled "RFI Reduction." A 0.125 aluminum chassis is planned. The upper dashed line represents the broadband spectrum of the 320-volt rfi. Frequencies below 14 kc are not applicable per Mil-I-11748. Thus, the first applicable frequency is the worst case frequency. As can be seen, the 0.125 chassis provides adequate attenuation.

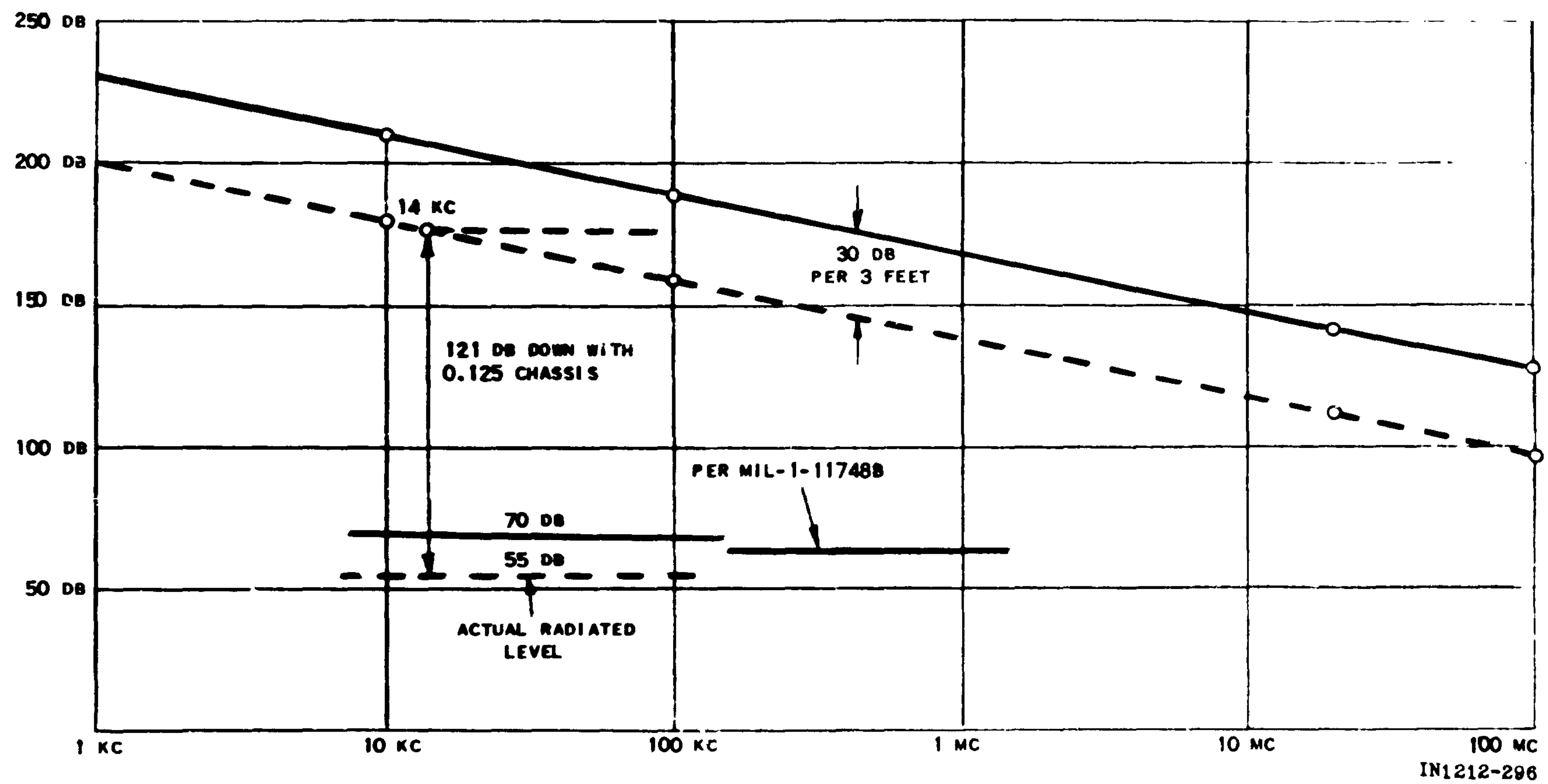


Figure 1-29. RFI Reduction -320 Volt, 400 CPS Square Wave, 0.125 Al. Chassis

(2) In the event that some forced air cooling is required, suitable screening will be utilized. It is anticipated that an air intake and exhaust would be rfi filtered with honeycombed aluminum which would provide effective shielding of 70 - 100 db.

(3) The access panel (top) will be rfi gasketed and attached with 3/4-turn fasteners spaced no more than two inches apart. Refer to figure 1-25. RFI gasketing attenuation is 97 db. 97 db attenuation will result in a db level of 76 db. Over 30 db of attenuation will be provided by the shelter and racks, giving an overall db level of 46 db maximum.

(4) Input leads will be kept a maximum distance away from the output leads to the blower.

(5) A large capacitor will be placed across input leads (dc).

(6) The output square waves will be Y-connected to a Y-connected motor. This will significantly reduce the harmonic content and virtually eliminate the third harmonic. In addition, these leads will be shielded.

(7) Input and output leads will enter through filter boxes which will provide rfi attenuation. These boxes with shielded leads provide adequate rfi suppression. This conducted noise decrement due to shielded cable is shown in figure 1-30. The actual level is significantly below the level stipulated in MIL-I-11748.

1-37. RFI Design Plan for AN/GRC-( ) Ancillary Equipment

a. General. Throughout this report under the mechanical considerations for each unit, the word phrase "containment" appears several times. The reader's attention is directed to figure 1-31 which indicates that the minimum attenuation afforded by an aluminum chassis 0.090 inches thick is in excess of 100 db at a frequency of 14 kc. It should be noted that the attenuation greatly increases with frequency. The chassis discussed herein will be constructed of 0.090 aluminum stock. Similarly, throughout this

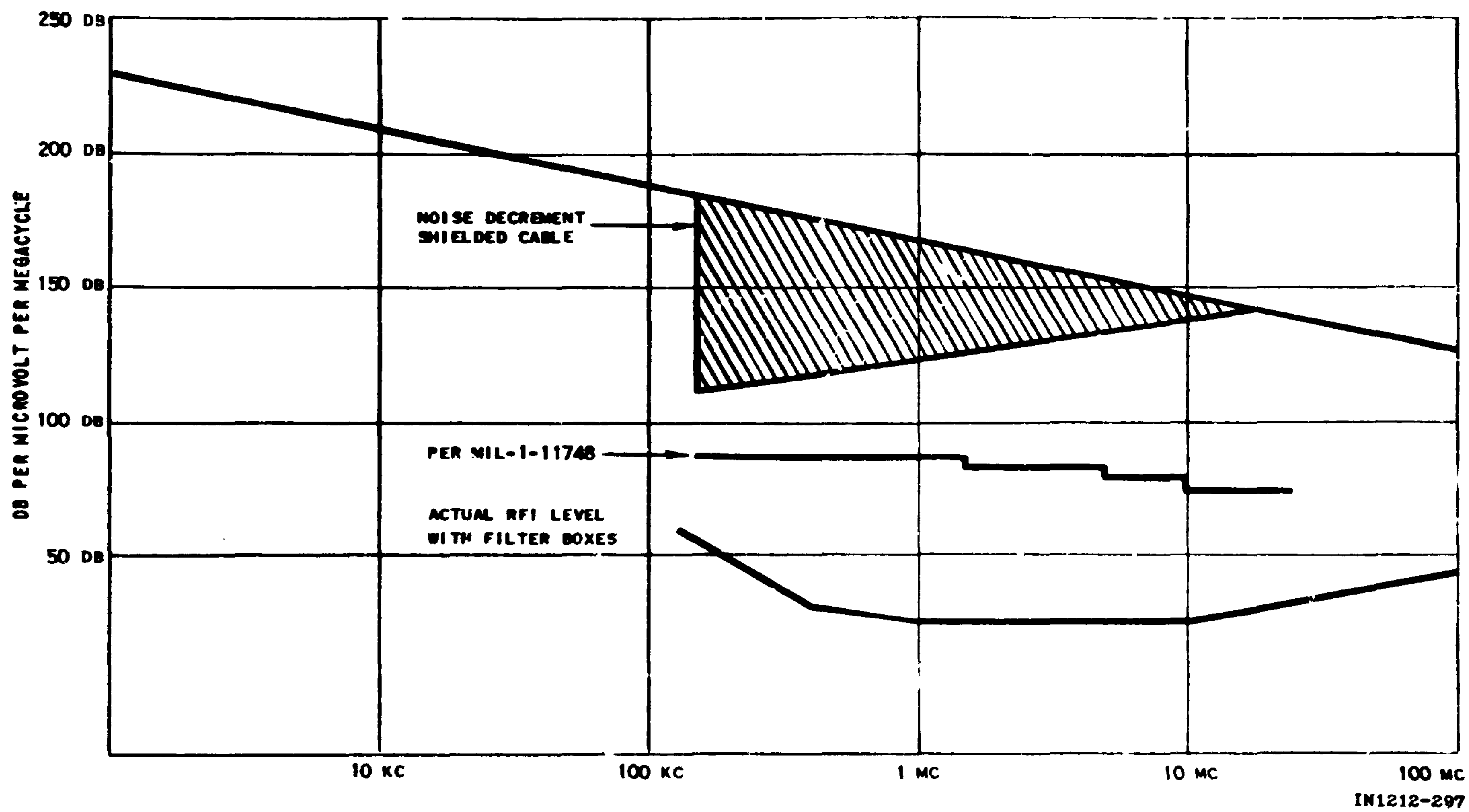


Figure 1-30. RFI Reduction -320 Volt, 400 CPS Square Wave, Shielded Cable

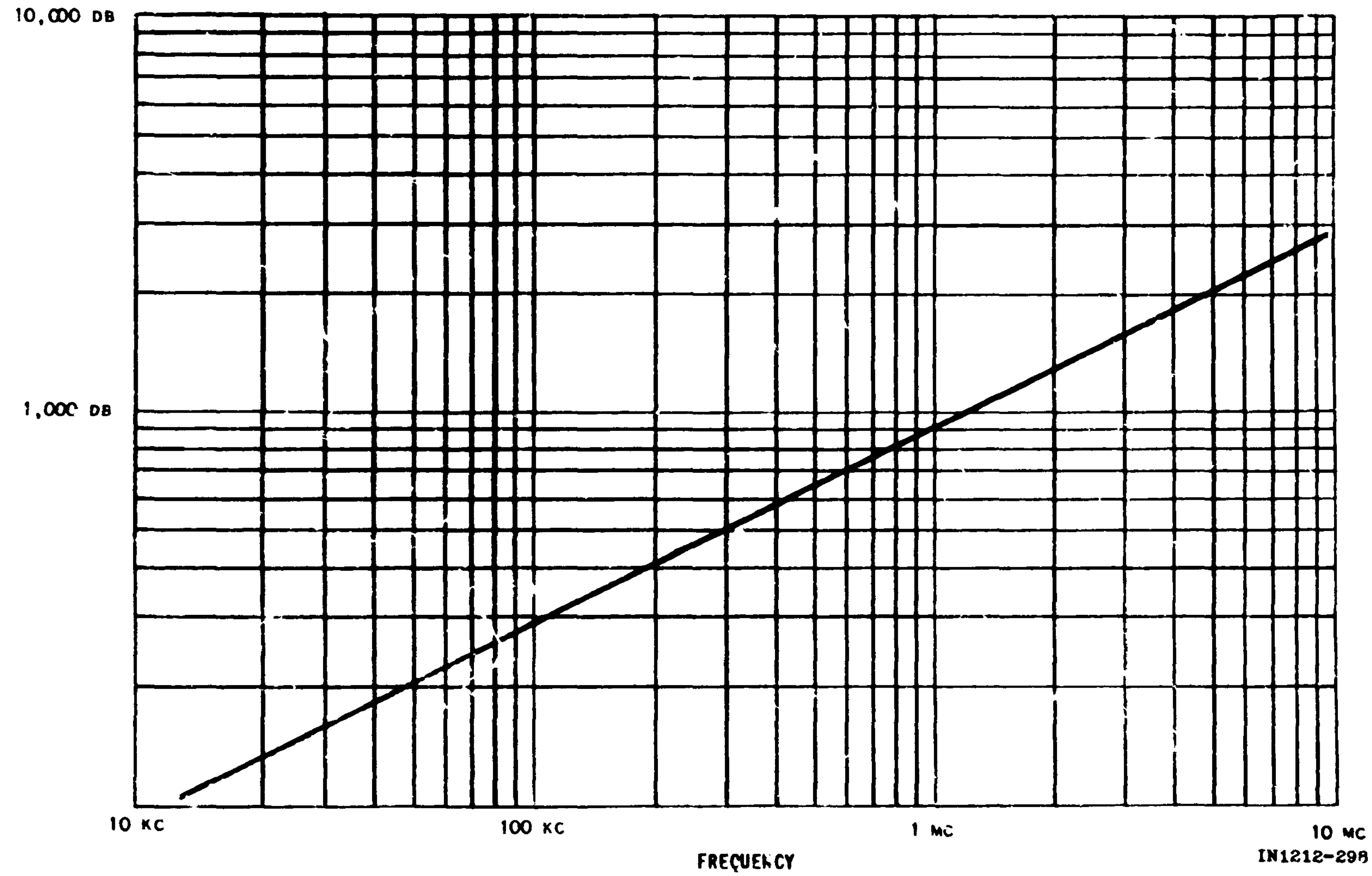


Figure 1-31. RF Attenuation vs Frequency Al Chassis .090 In. Thick

report, under mechanical considerations, it will be noted that a screw spacing of 1.75 inches has been used. According to figure 1-32, this screw spacing should provide about 80 db attenuation.

b. RFI Considerations.

(1) Local-remote switching

(a) Electrical considerations

| Interference Source | Fix |
| --- | --- |
| 28-volt dc power | The input lines will be filtered using separate rfi containers to house the filters. . filters will provide a minimum of 60 db attenuation to .15 mc signals, 90 db at .5 mc, 100 db at 1.0 mc, 100 db at 10 mc, and 94 db at 100 mc |
| Remote input from field lines | The field line inputs will be filtered with the same type filters as the input lines. Again, the filters will be mounted in separate rfi containers |
| Local TTY input and output lines | The local TTY input and output lines will be routed in the interference free portion of the switching panel case. Any interference radiated by the TTY lines will be contained by the .090-inch thick aluminum chassis. This thickness of aluminum should provide greater than 100 db at frequencies in excess of 15 kc (fig. 1-22) |

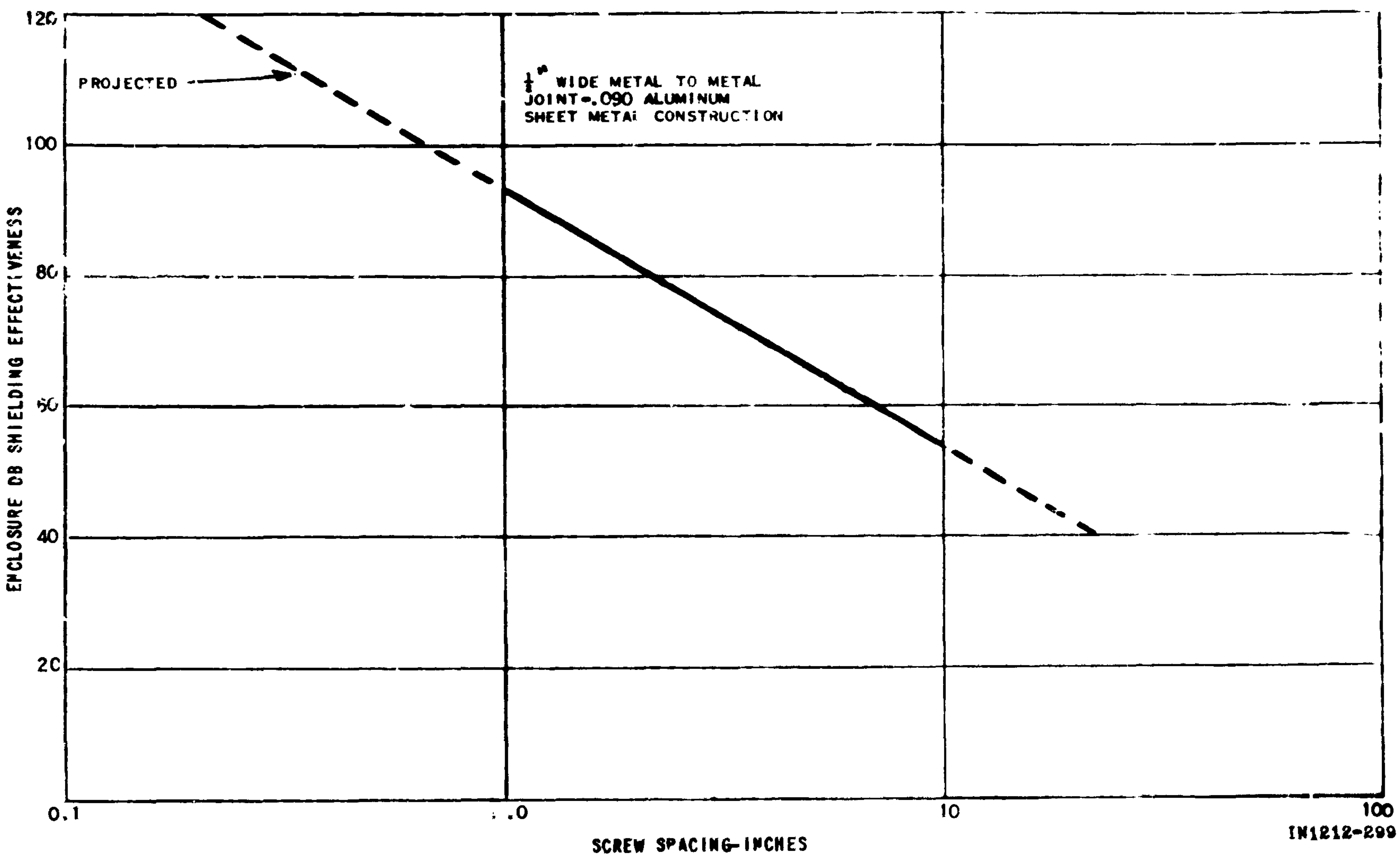


Figure 1-32. Shielding Effectiveness vs Screw Spacing

(b) Mechanical considerations

| Interference Source | Fix |
|---|---|
| Case | One piece, folded and welded |
| Front panel | One piece, 1.75" to 1.84" screw spacing. This screw spacing should afford over 80 db shielding |
| Top cover | One piece, 1.75" spacing between screws |
| Filter boxes | One piece construction, folded and welded |
| Connectors | MTD on paint-free iridited surface |

(2) Receiver dummy box

(a) Electrical

| | |
|---|---|
| Single pair (with TTY signals) enters and exits | Containment |

(b) Mechanical

| | |
|---|---|
| Case | One piece, folded and welded |
| Cover | One piece, 1.75" screw spacing |
| Connectors | MTD on paint-free iridited surface |

(3) Transmit dummy box

(a) Electrical considerations

| | |
|---|---|
| TTY signal lines entering 80 ma level | Containment |
| Polar relay following switching TTY signals | Containment |
| 28-volt dc bias used on polar relay coil | Containment |

| Interference Source | Fix |
|---|---|
| 28-volt dc interlock for TTY patch panel circuitry | Containment |

(b) Mechanical

| | |
|---|---|
| Case | One piece, folded and welded |
| Top cover | One piece, 1.75" screw spacing |
| Connectors | MTD on paint-free iridited surface |

(4) TTY patch

(a) Electrical

| | |
|---|---|
| Four relays, three of which are operated in conjunction with the duplex one-way switch and the Transmit/Receive switch. The 4th operates continuously with dummy boxes in place | Containment. The relay coil lines will be filtered at the unit doing the switching. The filters employed will have the same characteristics as the input line filters. Diodes will be placed across relay coils to suppress transients |
| TTY signal lines entering, being routed by above relays and exiting to dummy boxes | Containment |

(b) Mechanical

| | |
|---|---|
| Case | One piece, folded and welded |
| Top cover | One piece, 1.75" screw spacing |
| Connectors | MTD on paint-free iridited surface |

(5) Transmit/receive switch

(a) Electrical Considerations

| | |
|---|---|
| Switch S1. The switch applies power and | Filter all lines entering and leaving the box. The attenuation |

| Interference Source | Fix |
| --- | --- |
| switches relays located in the TTY patch panel. | characteristics of these filters have been discussed under "28-volt dc power" |
| (b) Mechanical considerations | |
| Case | One piece, folded and welded |
| Cover | One piece, 1.75" screw spacing |
| Connectors | MTD on paint-free iridited surface |
| (6) Power entrance panel | |
| (a) Electrical considerations | |
| Generator noise from truck can conceivably enter through J1 | Filter both input lines from the truck generating system. The filters will have the attenuation characteristics discussed |
| Ignition noise and other power equipment noise can conceivably enter through J2 | Filter both input lines from the power source. The filters will have the attenuation characteristics discussed |
| TTY land lines and the local/remote unit land lines may carry rfi into the shelter | Feed-through capacitors will be used to prevent rf from entering the shelter on TTY lines. It is necessary to preserve up to about the 20th harmonic of the fundamental TTY signal; and hence feed-through capacitors will be used |
| (b) Mechanical | |
| Case | Five formed metal parts welded to form a continuous surface with no discontinuities |

| Interference Source | Fix |
|---|---|
| Cover | One piece construction with 1.75" screw spacing |
| Connectors | MTD on paint-free iridited surface |

(7) Power distribution panel

(a) Electrical considerations

| | |
|---|---|
| Air conditioner | The lines to the air conditioner will be filtered at the power distribution panel using two filters which provide 60 db minimum at 150 mc, and 80 db minimum 1.0 mc to 1000 mc |
| Prime power converter | The lines to the converter will be filtered at the power distribution panel using filters which provide 50 db at .15 mc, 90 db at 1.0 mc, 90 db at 10 mc. It should be pointed out that filtering is in addition to the filtering done at the prime power supply itself |
| TTY power converters | The ac lines to the TTY converters will be filtered using filters having the characteristics discussed under "28-volt dc power". The emergency 28 vdc power lines will be filtered using filters of the characteristics discussed. |
| Security equipment | No filtering will be used |
| RT-( ) Transceivers | The power lines to transceivers will be filtered using filters of the characteristics discussed. |

| Interference Source | Fix |
|---|---|
| Modem | The power lines to the modem unit will be using filters with the same attenuation as input line filters |
| Local/Remote Switching Panel | Lines at the L/R panel are filtered in the L/R panel |

(b) Mechanical considerations

| | |
|---|---|
| Case | The chassis case will be folded and welded construction |
| Cover | The one piece cover will have 1.75" screw spacing |
| Front Panel | The speaker holes will be covered by perforated metal. The perforated metal will consist of .0625" dia. holes on .125" centers. This should afford approx. 60 db of shielding effectiveness. The front panel will have 1.75" screw spacing. All metal to metal contacts will be paint-free, iridited surfaces |
| Connectors | The connectors will be mounted on a paint-free, iridited surface |
| Potentiometers | The speaker volume control potentiometers will employ rf gasketing material to reduce rfi. The rf gasket, in the manner used, will afford a shielding effectiveness of from 20 db to 40 db at 1000 mc |

(8) TTY power converter

| Interference Source | Fix |
|---|---|
| AC to DC and DC to AC Converter | The power lines to these units will be filtered as the input lines. While little control can be exercise over the radiated interference of the interim units, the militarized units will certainly meet Mil-I-11 |

(9) Shelter accessories

| | |
|---|---|
| Bell | The doorbell will be housed in a perforated metal enclosure affording rfi suppression. The perforated metal will consist of .0625" dia. holes on .125" centers. This should afford approx. 60 db of shielding effectiveness |
| Buzzer | The buzzer will be housed in a perforated metal enclosure affording rfi suppression. The perforated metal will consist of .0625" dia. holes on .125" centers. This should afford approx. 60 db of shielding effectiveness<br>Filters may be used in the power lines to the buzzer |
| Door Viewer (to inspect personnel requesting admittance to the shelter) | The door viewer is 2.1 inches long with a .350" inside diameter. Based on the approx. formula: db = 32 L/d, this unit should afford in excess of 100 db attenuation |

| Interference Source | Fix |
|---|---|
| Mail slot | The interior side of the mail slot will have a spring-loaded flap; and the exterior side of the mail slot will have a watertight spring-loaded flap with quick release fasteners. This will afford virtually the same rfi integrity as the continuous shelter skin |

(10) Local remote unit

(a) Local unit

1. Electrical considerations

| | |
|---|---|
| RF introduced on incoming audio land lines and key lines from the remote unit | Filter all incoming lines using feed-through capacitors. The feed-through capacitors are suited for 2 - 30 mc interference reduction. A 20 to 35 db attenuation can be realized |
| Keying relay | The placement of a diode across the relay coil will suppress transients |
| Pickup from unshielded handset cord introduced into the unit | The incoming handset leads will be filtered with feed-through capacitors |

2. Mechanical considerations

| | |
|---|---|
| Case | The case will be a continuous container made by folding and welding a single piece of metal |
| Front Panel | The front panel will be single, heavy cast aluminum piece |

(b) Remote unit

| Interference Source | Fix |
|---|---|
| 1. Susceptibility considerations | |
| RF introduced on incoming audio land lines and the interconnecting key line | All incoming lines will be filtered by using feed-through capacitors. The feed-through capacitors will provide 20 to 35 db attenuation over the 2 to 30 mc band. |
| RF introduced by the use of a handset with an unshielded cord | Feed-through capacitors will be used to filter the incoming handset lines |
| 2. Mechanical considerations | |
| Case | The case will be a container made from a single folded and welded piece of metal |
| Front Panel | The front panel will be a single, heavy cast aluminum piece |

c. Cabling and wiring of the shelter

(1) Method for cable construction

(a) Shielding

1. All cables, both power and signal, will be double shielded. The shield will not be used to carry any signal or power return, or any current.

(b) Grounding

1. Where applicable, grounding washer with evenly distributed, flared, soldered shield method will be used.

2. In cases where the grounding washer cannot be used, the shield will be carried by a pin through the connector to ground.

(d) Insulation

1. The shielded cable will have an overall insulating sheath

(2) Shelter ground system

(a) Bus bar

1. A bus bar, consisting of a piece of copper bar, 3/8" by 3/4" in cross section, will be used for a common ground for all units in the shelter

2. The bus bar will be insulated from the shelter skin by use of standoffs. Only one point will be grounded. This point will be the grounding stud facility which protrudes to the outside

(b) Ground stake

1. The ground stake will be five feet in length

2. A six-foot ground strap will be fixed to the ground stake for connection to the shelter ground stud

(3) Shelter racks

(a) The shelter racks should provide approximately 5 to 10 db of attenuation

(b) Filters

1. The shelter rack air ducts shall incorporate wave guide below cutoff filters at the intake and exhaust ports. These filters will provide a shielding effectiveness of at least 70 db for all frequencies below cutoff, which will be 24 gigacycles

(4) Shelter

(a) The shelter door shall have compressible rfi gasketing around its perimeter. This gasketing will be bonded to, and will mate to paint-free iridited surfaces

(b) It is anticipated that the shelter enclosure itself will provide approximately 20 db attenuation.

1-38. RFI Design Plan for Government Furnished and Contractor Purchased Items

The status of the above items regarding their rfi capability has been advised in USAELRDL letter SELRA/( )- specify contract number__ __ ______. dated ________________.

a. S-( ) Shelter - The shelter has not specifically been designed to meet Mil-I-11748B. Tests to determine the rf shielding integrity have only been made from 26 mc up. Attenuations were never less than 30 db. For the purpose of the AN/GRC-( ), we are considering 20 db in the 2 - 30 mc range to be a typical figure.

b. AN/TGC-( ) - Complies with Mil-I-16910 (Navy); does not comply with Mil-I-11748B Case and Cable Radiation, Class III a requirements, in the frequency range of 14 to 150 kc. Fails Mil-I-11748B by 32 db at 14 kc.

c. AN/TGC-( ) - No rfi data available, was not evaluated by USAELRDL. Same as AN/TGC-( )except for additional features.

d. GEH-9-A-60 - Not tested for rfi. (A similar version, the CE-6-A-60, failed rfi tests.)

e. M-37 - Engine and charging system comply with Mil-I-10379. Due to the delay in receiving the information, a comparison has yet to be made with Mil-I-11748B.

f. Trailer Generator - Engine is a MIL-STD Model 4A084 and meets Mil-I-11683B. Alternator is a ________________ Electric Company Model 5660 and will be manufactured to meet Mil-I-11683B.

g. H-33F-PT - No rfi data available - may be susceptible to 0.1 v/m rf fields as required by Mil-I-11748B.

h. M-29/U - No rfi data available - may be susceptible to H fields in the audio range and to 0.1 v/m rf field as required by Mil-I-11748B.

i. RT-( ) - The RT-( ) is designed to meet SCL-( ). This specification calls for Mil-I-11748A (not B) and only in part. For example, there is no susceptibility requirement.

## Section VI. ELECTROMAGNETIC COMPATIBILITY TEST PLAN

### 1-39. EMC Test Plan

The following Electromagnetic Compatibility Test Plan is a typical example of a good emc test plan submitted to the Department of the Army by an electronic manufacturer.

Date

Report No.

ELECTROMAGNETIC COMPATIBILITY
TEST PLAN

FOR SER. NO.

RFI Specification MIL-I-

Issue DATED

Equipment Specification

Issue DATED

Contract No

**TESTS TO BE PERFORMED BY**

**Prime Contractor**

**RFI Test Laboratory**

**NOTE - If rfi tests are to be performed by a commercial testing organization, a complete brochure describing that company's test facilities and capabilities must be attached to this document.**

Figure 1-33. Electromagnetic Compatibility Test Plan Cover Sheet

Date
Report No.

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1. GENERAL

1.1 Scope

This procedure specifies the tests to be made to determine compliance of the equipment or system under test to Military Specifications on radio frequency interference.

1.2 Reference Specification

a. Specification MIL-I-__________(Class ☐ I ☐ II ☐ III ☐ Other, equipment) shall be adhered to in all tests performed.

b. Specification SCR ____________________ is the governing technical requirement for equipment under test.

c. Other________________________________________________

1.3 Data Sheets

a. All data will be recorded on Data Sheets.

b. Sample Data Sheets are included in this procedure and are pages ____________________of the Data Sheet section of this procedure.

c. An Equipment Log and Operational Log are also included and are pages ___________and____________ of the Data Sheet section of this procedure.

d. Appropriate Data will be plotted graphically. These graphs will be made to resemble those of Figures__________of MIL-I-__________, or as shown in Figure ______________ of this procedure.

e. For any system utilizing more than one receiver and transmitter, or any combination thereof, a radio frequency spectrum chart shall be prepared and become part of the test plan. See appendix for suggested method.

2. TEST EQUIPMENT AND FACILITIES REQUIRED

2.1 Test Area

All tests in this procedure shall be performed in an interference free test area. It is desirable that the ambient interference level 1.

during these tests, measured with the test sample de-energized, be at least 6 db below the allowable specification limit. However, in the event that at the time of measurement the levels of ambient interference plus test sample interference are not above the specification limit, the test sample shall be considered to have met the specified requirements. All non-radiation type tests may be performed in a shielded enclosure.

2.2 Test Equipment (Noise and Field Intensity Meters)

| Manufacturer & Model No. | Series Nos. (one each) | Last Date of Calibration |
|---|---|---|
| List all noise and field intensity meters and associated antennas. | | |

2.3 Miscellaneous Test Equipment (List if Required)

a. Audio Oscillator, 50 to 20,000 cycles. Indicate output impedance and level.

b. Audio amplifier capable of delivering 3 volts rms into 5 ohms over a range of 50 to 20,000 cycles (35 to 50 watts).

c. AC Voltmeter, Range____________________.

d. Audio induction probe____________________.

e. RF signal generators to cover the range______kc to ______kmc. with an output level of__________microvolts and output impedance of 50 ohms.

f. Low, high, band pass and band rejection filters as required.

g. Directional couplers, rf samplers and attenuators as required.

h. Line impedance stabilization networks necessary to comply with specification (see MIL-I-______par.__________).

i. Monitoring oscilloscope, plotter, x/ or linear.

3. TEST SETUP AND CALIBRATION OF TEST EQUIPMENT

3.1 Test Specimen

a. The general arrangement of the test sample and interconnecting cables shall simulate actual installation insofar as practical.

The position on the ground plane shall be indicated in figure ________ of this procedure. Where equipment size exceeds ground plane dimensions, or when more than two line impedance stabilization networks are used, the arrangement shall be as close as possible to above.

b. The test sample modes of operation and control settings shall be chosen as those expected to produce maximum interference or susceptibility. Specify modes of operation in Table ________ of this procedure.

c. Any dummy antenna used shall simulate as closely as possible the electrical characteristics of the system antenna, and shall be capable of handling the system output power.

3.2 Test Equipment

a. The antenna of the radio interference meter shall be positioned as directed in Figure__________ of MIL-I-_________________.

b. The detector function to be used on all broad-band and pulsed cw interference will be PEAK for the NF-205, NF 112-FIM and CFI. CW interference will be measured with the NF-205 in the CARRIER position, and the FIM in the AVERAGE position.

c. All test equipment shall be periodically checked and calibrated in accordance with manufacturer's instruction material.

d. Above 21 gc, Polarad Model "R" receiver will be employed. Specific instructions for its use will be included as part of this procedure.

4. TEST OPERATION

4.1 Conducted RF Interference (All classes of equipment)

a. With the test sample connected, figure _________, connect the Empire Devices NF-205 to the 50-ohm output of the stabilization network in the line indicated in the table below. Operate the system in Mode_____________ Table No._________ indicates control settings for each mode, scan the frequency range 150 kc to 65 mc. and enter the results on Data Sheet. The 30 ft. antenna cable supplied with the NF-205 shall be used to connect the noise meter to the line stabilization network. All unused stabilization networks shall be terminated in 50 ohms.

Stabilization Network in Line

| Network No. | Line Function | Network Code No. |
|---|---|---|

b. Repeat 4.1.a. with the system operating in (all other specified modes).

c. This data for each line shall be plotted on figures__________ MIL-I-___________ to determine acceptability by comparison with the "Limits" tables and/or curves.

4.2 Radiated Interference (All classes of equipment)

a. With the system connected, figure__________, connect the appropriate antenna to the interference measuring instrument. Operate the system in Mode______ and place antenna in specified position (figure _______ of this procedure). Scan the frequency range ________kc to ________kmc, and enter results on Data Sheet.

b. The interference measuring instrument shall be slowly tuned through each frequency octave and at least three frequencies per octave, at which maximum interference either cw or broadband is obtained, shall be selected as test frequencies. (Note on Data Sheet) In addition, particular notation shall be made of cw interference levels at the following frequencies and their harmonics:

(Those frequencies and their tolerances which are unique to the system under t.st. They are dependent upon circuits within the system having a tendency to radiate, such as frequency multipliers, crystal saver circuits, beat frequency oscillators, local osciiiator frequencies and harmonics.)

This data shall be plotted on figure _______ MIL-I-__________ to determine acceptability by comparison with the "Limits."

c. Repeat 4.2.a. with appropriate antenna in (all other specified positions).

d. Repeat 4.2.a and 4.2.b with system operating in (all other modes).

4.3 RF Radiated Susceptibility Class Ia and IIIa

a. Connect the rf signal generator and test antenna as shown in figure_______ MIL-I-________. Place the test antenna in specified position (figure _________ this procedure). Operate the system in Mode_______________. With specified input to the test antenna, and producing 0.1v/m field intensity as measured at the receiving antenna and at the distance specified below, direct the rf energy in turn to every face of the test specimen. Modulate the rf signal source with signals which develop the maximum susceptibility of the test specimen. Complete table below as required. No change in indication, malfunction or degradation of performance shall be produced in any equipment. Define degradation of performance. Enter observations on Data Sheet.

| Source Output | Frequency | Type Mod | Antenna Type | Test Distance |
|---|---|---|---|---|

b. Repeat 4.3.a with system operating in (all other specified modes).

4.4 RF Conducted Susceptibility (power or non signal lines) Class Ia and IIIa

a. Connect the rf signal generator to the rf connector of stabilization network in _____ power input lead. The output of the signal generator shall be 100,000 microvolts rms with specified modulation. Operate the system in Mode_________. No change in indication, malfunction or degradation of performance of the system shall be produced as the signal generator is tuned through the range 150 kc to 65 mc. Note on Data Sheet any observations; also record the output level of the susceptible signal.

b. Repeat 4.4.a. for (all other specified power input leads).

c. Repeat 4.4.a. and 5.4.b. for (all other specified system modes of operation).

4.5 AF Conducted Susceptibility (Class Ia and IIIa)

a. Apply sine-wave signal of __________v to _______v rms into ____ power input lead of system as shown in figure__________. Operate the system in Mode__________. No change in indication, malfunction or degradation of performance shall be produced as the frequency of the signal is varied from 50 - 20,000 cycles. Note on Data Sheet any observations. Also record output level of susceptible signal.

b. Repeat 4.5.a. for (all other specified power input leads).

c. Repeat 4.5.a and 4.5.b. for (all other specified system modes of operation).

4.6 Front End Rejection (Receivers only)

a. Connect the signal generator as shown in figure ________ to the input of the receiver of the system under test with a 50-ohm cable. The signal generator output, when tuned to the receiver frequency, which gives the minimum perceptible reading above receiver background noise, shall be noted at each end and center of the tunable range of the receiver. (The signal generator may be modulated for convenience.) Operate the system in Mode_________ and record on Data Sheet as V., readings taken in all modes for the three frequencies specified above. Perceptible signal above receiver background noise may be established as $\frac{S + N}{N}$ = 10 db

$$\text{Front End Rejection} = 20 \log \frac{V_2}{V_1}$$

$V_1$ = Signal generator voltage required for minimum perceptible receiver output at frequency under test.

$V_2$ = Signal generator voltage required for minimum perceptible receiver output at all other frequencies.

(Note: If appropriate filter networks are not available, allow for signal generator harmonics when evaluating response at points away from receiver setting.)

b. With the receiver tuned to the low end of the tunable range and the system in specified mode, note signal generator output necessary to give minimum perceptible receiver output from ________ kc to _______ kmc. Enter on Data Sheet as $V_2$ the output at each octave, except near receiver tuned frequency, where readings shall be taken often enough to permit preparation of an accurate response curve.

c. Repeat 4.6.b. with system operating in (all other specified modes).

d. Repeat 4.6.b. and 4.6.c. for (center frequency).

e. Repeat 4.6.b. and 4.6.c. for (high end of tunable range).

f. This data for Front End Rejection shall be plotted on graph paper to determine acceptability by comparison to the "Limits" of MIL-I-_________.

7.

4.7 Intermodulation (Narrowband)

a. Operate the system in Mode_____________, and connect two rf signal generators simultaneously to the antenna input of the system receiver as shown in figure __________. Adjust the output of each signal generator to 100 db above 1 microvolt at the antenna input. Modulate one signal generator 30% at 400 cycles, and the other 30% at 1,000 cycles. Tune one signal generator to a frequency above that of the receiver ($f_1 = f_{rec} + f_x$), and the other to a frequency lower ($f_2 = f_{rec} - f_y$). These frequencies shall be chosen such that their sum or difference equals the receiver frequency, and neither will give an output when applied alone.

b. Repeat 4.7.a. with the system operating in (all other specified modes).

c. Repeat 4.7.a. and 4.7.b. above, with the receiver connected into the system, and the rf signals fed into the directional coupler.

4.8 Intermodulation (Broadband)

a. Operate the system in Mode__________ and connect the standard impulse generator to the receiver input by means of a 30-ohm cable terminated with a 10-db pad. Adjust the output attenuator of the impulse generator for minimum perceptible receiver output. Record impulse generator output in dbmc.

b. Measure and record receiver local oscillator (lo) voltage at the input to the mixer circuit. Use a high input impedance device of at least several megohms or greater. If the test specimen is of a multiple conversion type, then measure and record each lo voltage.

c. Disable each lo and inject into the mixer circuit a 60 cycle voltage or current equal to the values obtained in b. above.

d. After determining that the receiver is functioning properly, increase the output of the impulse generator until the minimum perceptible receiver output is again obtained. Record the impulse generator output in dbmc. This generator setting in db, less the original setting in db, is the receiver broadband rejection in db. Compare results with MIL-I-_______ requirements.

4.9 Antenna Conducted (Class I equipment only)

4.9.1 Key-Up

a. Operate the system in Mode_______. Connect the interferencemeasuring instrument to the system antenna (transmitter output) as shown in figure ________ and scan the frequency range ___________kc to _____________kmc, for any rf output. At least 3 readings per octave shall be taken; and also the nature of the output (i.e., broadband or cw) shall be recorded on Data Sheet.

b. Repeat 4.9.1 a. with system operating in (all other specified modes).

c. Repeat 4.9.1. a and 4.9.1, b. with interference measuring instrument connected to receiver input terminals. If common antenna terminals are used for both trans- and receiver, operate receiver with trans- shut off.

d. This data shall be compared to the limits of MIL-I-___________ to determine acceptability.

4.9.2 Key Down

a. See figures______________ for test set-ups. The transmitter shall be operated into a dummy load. A suitable coupling device shall be used to sample the transmitter output and protect the measuring equipment. Attention should be given to oscillator frequency and harmonics, outputs from frequency multipliers and crystal saver circuits, beat frequency oscillators, etc. Spurious emanations shall be below the fundamental by the value specified in MIL-I-__________ throughout the frequency range. __________kc to __________kmc.

b. Insert suitable coupling device in series with the antenna line of the system and the rfi measuring equipment. Measure the fundamental power, frequency and record on Data Sheet. Fundamental power may be expressed as db/1µv for 500 kw systems.

c. Connect an appropriate spectrum analyzer to the directional coupler output or sampler device and photograph output spectra at the fundamental frequency. Compare data to equipment or system specification requirements.

10.

1-40. Appendix to EMC Test Plan

The following appendices and sample data sheets are part of the EMC Test Plan.

APPENDIX A

TEST PROCEDURE

MODES OF EQUIPMENT OPERATION

Operating mode of system under test vs. test

4.1 Conducted Interference

4.2 Radiated Interference

4.3 RF Radiated Susceptibility

4.4 RF Conducted Susceptibility

4.5 AF Conducted Susceptibility

4.6 Front End Rejection

4.7 Intermodulation Narrowband

4.8 Intermodulation Broadband

4.9 Antenna Conducted

4.9.1 Key Up

4.9.2 Key Down

Note: For each test indicated above specify methods of loading, triggering, operation of, and control settings on, test sample which will produce maximum interference and susceptibility.

11.

APPENDIX B

TEST REPORT

A test report conforming to Specification MIL-T-9107 shall be submitted to the procuring activity prior to submission of the development model for acceptance. In addition to the requirements in Specification MIL-T9107, the test report shall include the following details of testing:

a. Approved test plan as referenced herein.

b. Nomenclature of interference measuring equipment.

c. Date of last calibration of interference measuring equipment.

d. Detector functions used on interference measuring equipment.

e. Internal noise level of instrument at each detector function used at each test frequency.

f. Descriptions of procedures used.

g. Measured line voltages to test sample.

h. Test frequencies.

i. Method of selection of test frequencies.

j. Type of interference measured.

k. Measured level of interference and susceptibility at each test frequency.

l. Specification limit at each test frequency.

m. Graphs showing items (e), (h), (k) and (l).

n. Photographs of test set ups and test sample.

o. Sample calculations (showing how item (k) was obtained for each antenna used).

p. Screened inclosure, size and test data indicating compliance with MIL-STD-285.

q. Description and ambient profile data of interference free area.

12.

r. Measured impedance of line stabilization network.

s. The test sample shall be completely identified in the test report with complete nomenclature, manufacturer, and serial number. All suppression work performed on the test sample during the interference tests shall be fully described in words as well as by the test data in the report.

13.

EQUIPMENT LIST
RADIO INTERFERENCE TESTS

DATA SHEET NO. ______

| TEST PERFORMED | EQUIPMENT USED | SERIAL NO. | CALIBRATION DATA |
|---|---|---|---|
| | | | |
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Figure 1-34. Equipment List

14.

SUSCEPTIBILITY DATA
SHEET NO.______

EQUIPMENT UNDER TEST________________
SERIAL NUMBER________________
MODE OF OPERATION________________
SIGNAL GENERATOR USED________________
SERIAL NUMBER________________
FREQUENCY RANGE________________

TEST DATE________________
OPERATOR________________
WITNESS________________
TYPE OF TEST
RADIATED ☐
CONDUCTED ☐
LINE__________
TEST CONDITION________________

| TEST FREQUENCIES | | THRESHOLD SIGNAL LEVELS | OUTPUT LEVEL | SPECIFICATION SIGNAL LEVELS | DESCRIPTION OF RESPONSE |
|---|---|---|---|---|---|
| $F_1$ | $F_2$ | | | | |
| | | | | | |
| | | | | | |
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| | | | | | |
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| | | | | | |
| | | | | | |

REFERENCE:
FIG. NO.__________

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Figure 1-35. Susceptibility Data Sheet

15.

INTERFERENCE DATA
SHEET NO.______

EQUIPMENT UNDER TEST______
SERIAL NUMBER______
MODE OF OPERATION______
TEST SET USED______
SERIAL NUMBER______
MEASUREMENT TECHNIQUE______
DETECTOR FUNCTION______
FREQUENCY RANGE______
1 CORRECTION FACTOR =
2 SPECIFICATION

TEST DATE______
OPERATOR______
WITNESS______
TYPE OF TEST______
RADIATED ☐
CONDUCTED ☐
LINE______
TEST CONDITION______

| TEST FREQ. MC | METER READING DB | 1 CORRECTION FACTOR DB | FINAL READING DB | | 2 SPEC LIMIT DB | REMARKS |
|---|---|---|---|---|---|---|
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REFERENCE:
FIG. NO.______

IN1212-303

Figure 1-36. Interference Data Sheet

16

# CHAPTER 2

## GROUNDING, BONDING AND SHIELDING DESIGN THEORY AND PRACTICE

## Section I. INTRODUCTION

### 2-1. Interference Control Philosophy

There are two approaches to the reduction of interference: initial design for optimum interference reduction (from electronic equipment design through final production) and application of remedial interference control measures after equipment has become operational. Of these approaches, the former is preferred. The initial design approach entails early determination of interference generation and susceptibility characteristics of a particular piece of equipment in its operational environment, based upon equipment function, configuration, and performance and interference specifications. The equipment is then designed to meet both these performance and interference requirements. When equipment is in the breadboard stage, interference analysis and tests can be performed, and a preliminary interference and compatibility evaluation made. From such an evaluation, a realistic determination can be made of whether additional interference reduction design measures are needed. Following any modifications, equipment is retested, and, if acceptable, released for production.

### 2-2. Interference Reduction Design Techniques

a. The primary method of control is efficient circuit design, maximizing the energy in intelligence-bearing signals and minimizing spurious energy. All technical characteristics of the device must be considered. Shielding, filtering, bonding, and isolation of the interference-producing unit are means useful in the circuit design. To maintain shielding integrity, all lines entering or leaving enclosures must be decoupled by suitable filters. These may consist of simple feedthrough capacitors or more complex inductance-capacitance networks. Induction and radiation fields can be contained within equipment enclosures by having equipment cases of adequate thickness (with joints welded or otherwise continuous) and by having spring-contact fingers or conducting gaskets on the periphery of removable covers or doors. In the presence of strong radiation at a fixed frequency, such as microwave energy from a radar transmitter, tuned filters in the receiver antenna circuit are often quite effective. One of the best techniques for minimizing interference at low frequencies is to use bonding straps to connect points so that they are at the same potential. Such a bond prevents currents and spark discharges from one point to another. At microwave frequencies, however, bond straps begin to lose their effectiveness as suppression measures because bond straps have high impedance at these frequencies.

b. Interference control at the source is achieved by confining and dissipating interference energy being generated so that it cannot reach susceptible circuits or equipment by conduction, induction, or radiation. Designers should analyze causes of variations in electromotive forces and impedances, eliminate those that are not essential to proper operation of the equipment, and reduce essential ones to the absolute minimum. Some of the more typical categories of interference that can be suppressed at the source, and the standard methods for suppressing them are:

1) Interfering components -- replace with components that do not cause interference

2) Interfering components can be relocated to areas within an equipment where their interference effects are diminished

3) Interference components can be shielded, and lines entering the shield enclosure can be filtered

4) Interference coupling -- rewire device, where possible, to isolate interference carrying conductors from output conductors

5) Conductor radiation -- keep interference carrying conductors as short as possible

## Section II. GROUNDING

### 2-3. General

This section examines various grounding techniques that should be used by designers during the design of electrical and electronic equipment. During the design stages, establishment of definite and systematic cable, wiring, and equipment grounding procedures will ensure minimal interference difficulty after installation. It will also minimize the need for additional interference suppression after the equipment becomes operational. Although it is not possible to have a set of fixed rules governing the grounding of electronic or electrical circuits or equipment, the guidelines presented here can be adapted by the design engineer to particular grounding problems encountered.

a. Grounding Plane Requirements. A good, basic ground plane is the foundation for obtaining reliable, interference-free equipment operation. An ideal ground plane should be a zero-potential, zero-impedance body intertie system that can be used as a reference for all signals in the associated circuitry, and to which any undesirable signal can be transferred for its elimination. Ideally, it must be able to absorb all signals while remaining stable. An ideal ground plane would provide equipment with a common potential reference point anywhere in the system, so that no voltage would exist between any two points. However, because of the physical properties and characteristics of grounding materials, no ground plane is ideal, and some potential always exists between ground points in a system. A ground plane should be constructed of a low-impedance material, such as copper, and be large enough in length, width, and thickness to provide a minimum of impedance between its extremities at all frequencies. The ground plane for a fixed location plant facility should consist of a continuous sheet of expanded copper or a grid of copper conductors 10' x 10' OC spacing or less. The expanded copper sheet is recommended for multitransmitter plants. VLF transmitters present special problems frequently requiring unique solution. The ground plane should present the highest capacity possible to earth. It should extend continuously under all equipment areas in the building, under the footings, and extend six feet or more beyond the limits of the equipment area(s). High power multi-transmitter plants may require ground radials extending from the ground plane a quarter wave length at the lowest operating frequency to permit adequate decay of ground currents. The dc resistance to earth must be kept low to prevent large changes in ground plane potential produced by conducted or induced currents caused by lightning discharge. Ground stakes driven into the permanent water table and bonded to the ground plane will usually provide an earth connection of ten ohms or less. In extreme cases, where ground resistance is high and the permanent water table is deep, as in desert areas, drilled wells may be required to provide a low impedance dc ground. Equipment ground tie points may be provided by a conveniently located ground bus bonded to the ground plane by copper strap at intervals of six feet or less.

b. Grounding Techniques. There are three fundamental grounding techniques that can be employed. Figure 2-1 illustrates each of these techniques. They can be used separately or in combination. They are:

(1) Floating ground system. In the floating ground method, the ground plane is completely isolated from all circuits.

(2) Single-point grounding system. In the single-point grounding method, a single physical point in the circuitry is designated as a ground reference point. All ground connections are tied to this point.

(3) Multipoint grounding system. The multipoint grounding system is one in which a ground plane -- for example, an entire chassis -- is used instead of individual return wires for each of the circuits.

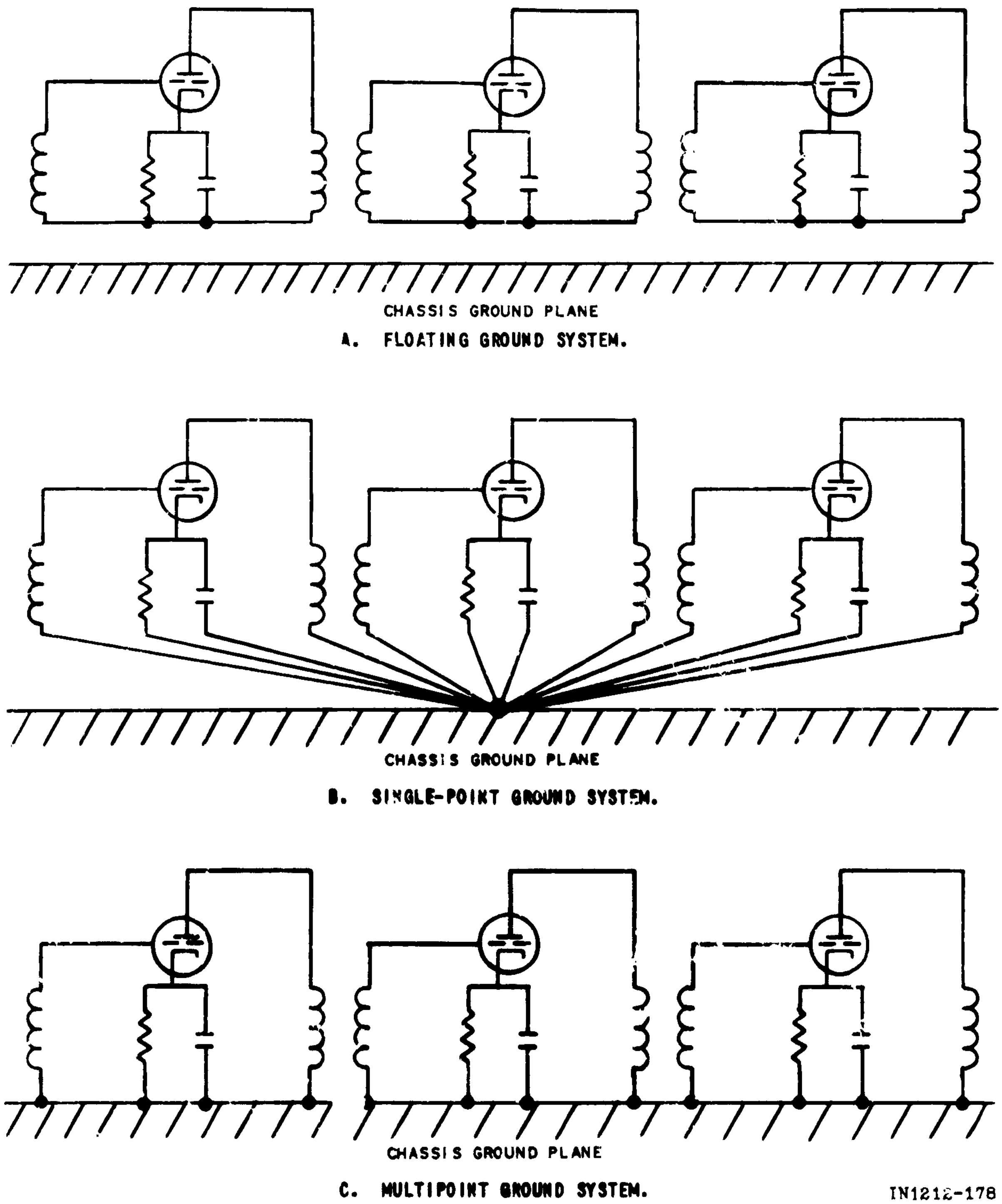



Figure 2-1. Ground Systems

c. Ground Connections. A ground point is the physical location where a circuit, piece of equipment, or system is connected to the ground plane. The impedance of a ground connection is a function of such factors as the size of conductors, the length of leads, and wiring techniques. If the ground connection is improperly made, it may be inadequate for satisfactory operation of the circuit and may, in fact, be more detrimental to the control of interference than if there were no ground connection. Grounding the rf portion of the spectrum is difficult and complex. Its complexity varies in direct proportion to the operating frequency. Several factors contribute to this:

1) Every wire has a definite inductance

2) A current flowing through a wire induces flux around the wire

3) As radio frequencies increase, inductive reactance causes circuit impedances to increase

4) The resonant frequencies of even small inductances acting with circuit capacitance often fall within the operating frequency of the circuit

5) As radio frequencies increase, skin effect becomes an important consideration

A low-impedance ground connection requires the ground leads to be as large and short as possible and to be securely bonded directly to the ground plane. A representative ground lug connection and its equivalent circuit is shown on figure 2-2. Figure 2-3A illustrates a typical method of connecting power and signal grounds. As the frequency increases, the inductance of the ground jumper can become appreciable and, if the power or signal circuit contains high-frequency interference currents, they may be conducted through the ground pin into the external wiring. In contrast, figure 2-3B shows the proper method of installing a ground to avoid conducting the interference through the connector.

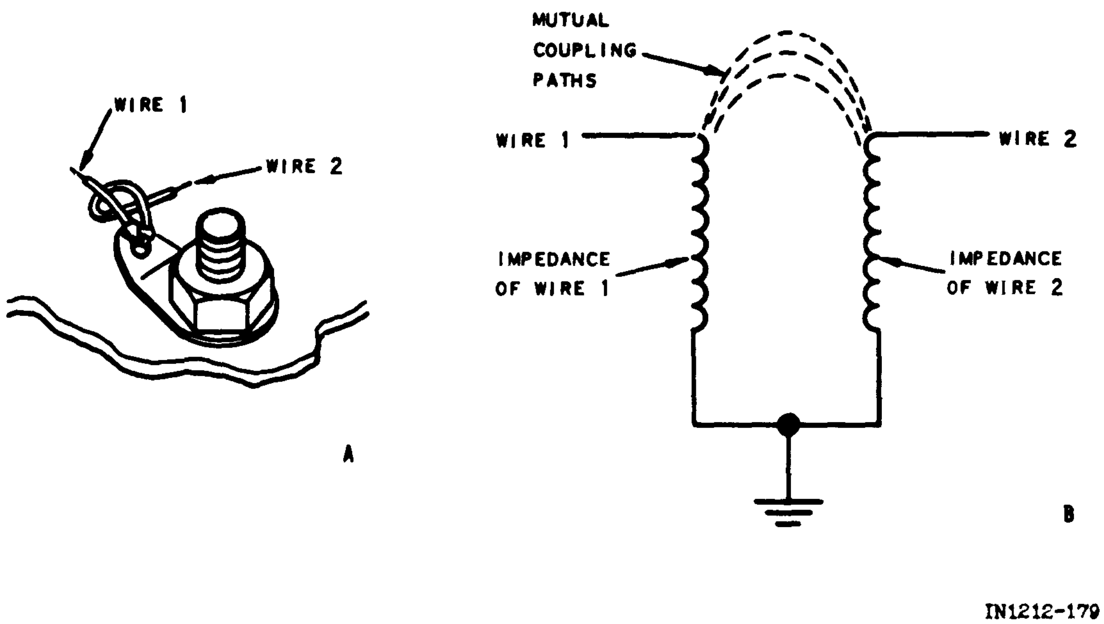
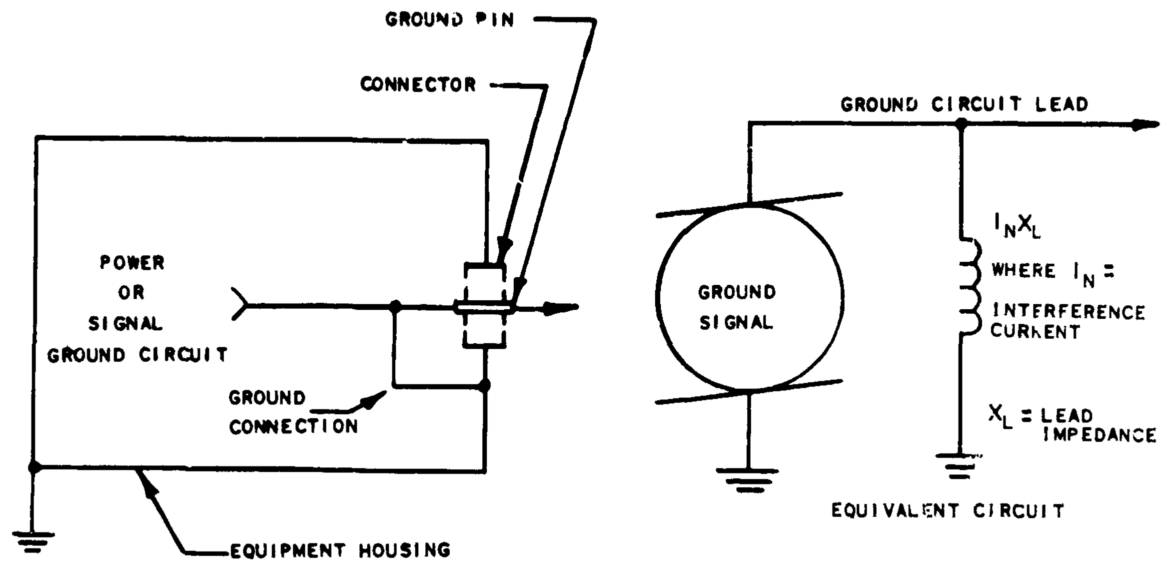


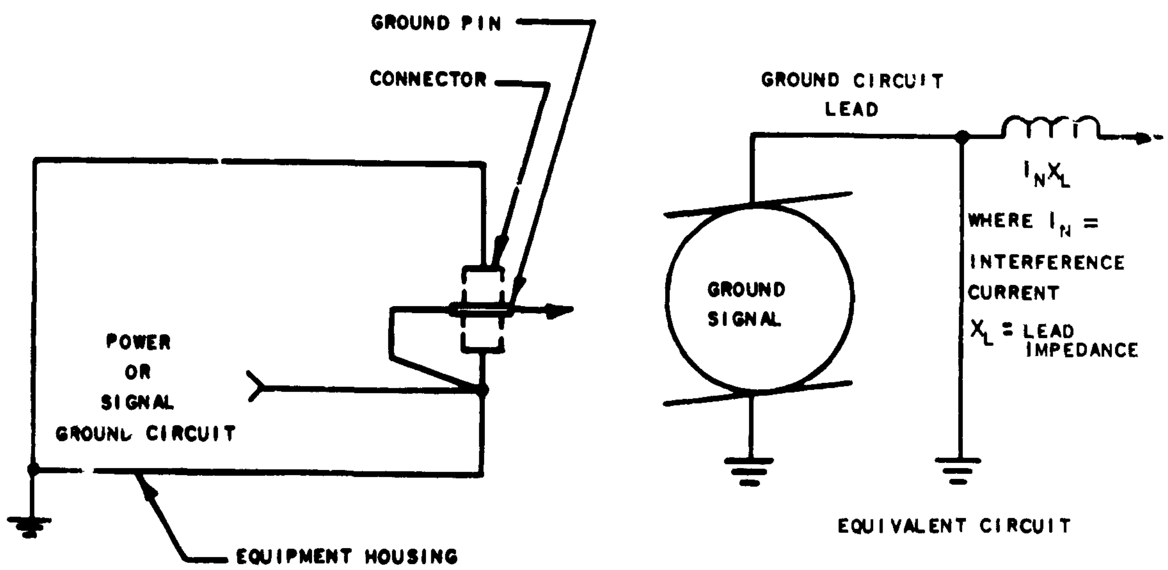

Figure 2-2. Ground Lug Connection and Equivalent Circuit

## 2-4. Distribution of Chassis Potential

A chassis is not always at zero-signal potential at all points. A typical plot of the chassis-potential of a ground plane is shown on figure 2-4. On this figure, the dark areas surround ground lugs or shields in high-voltage and/or high-current areas where the power is adequate to present a signal on the physical ground plane. Holes or other irregularities tend to increase the over-all resistivity of a ground plane. A potential plot, if available, would aid tremendously in determining the location of low potential or equipotential points that can be used in grounding small-signal-sensitive circuits such as grid bias resistors. A matter of an inch in the location of ground points can make a difference of several millivolts in the potential between two ground points. Often it becomes necessary for a designer, during the initial stages of development, to utilize a potential plot in selecting ground points. Even then, if an adequate but marginal ground-point location is selected, tolerances encountered in production can result in further difficulties so that a more discriminate choice of locations may be required. It may be advantageous at times to run a longer ground wire to utilize a cooler location on a ground plane.




A. METHOD COMMONLY EMPLOYED IN MAKING GROUND CONNECTIONS IN ELECTRICAL AND ELECTRONIC EQUIPMENT.




B. PROPER METHOD OF GROUNDING FOR MAXIMUM INTERFERENCE REDUCTION.

IN1212-180

Figure 2-3. Grounding Methods

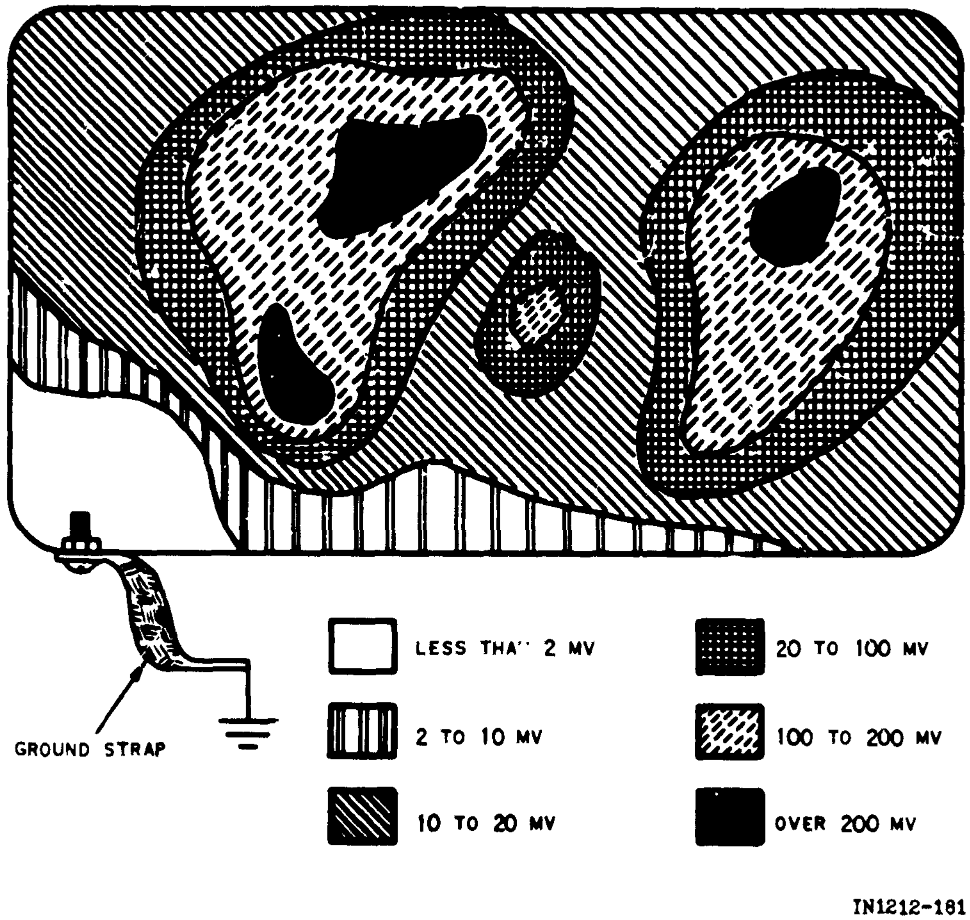



Figure 2-4. Typical Ground Potential Graph at a Specific Frequency

## 2-5. Shields

a. Equipment. Grounds for apparatus housed within a shield should be arranged so that the shield is not used as a return conductor. In this way, the current that flows through the shield is reduced to a very small value and the tendency for energy to leak out through holes or joints in the shield is minimized. The ideal grounding system makes use of a ground bus or ground plate within the shield that is insulated from the shield except at a single point (ground point). This ground point provides the only ground connection for the apparatus within the shield, as illustrated on figure 2-5.

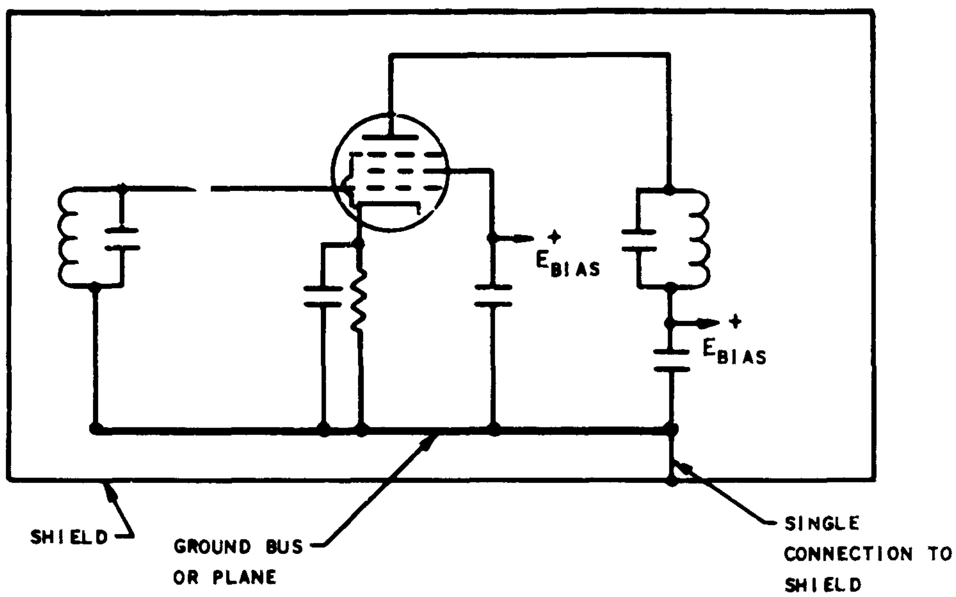



Figure 2-5. Single-Point Ground Bus Arrangement

b. Cable Grounding. The problem of electrical compatibility in a complex electrical or electronic system is in many cases dependent on the treatment of the shielding and the grounding of the shields of interconnecting leads. Injudicious application of a grounded shield to a wire may cause coupling problems that otherwise would not exist. Grounding of the shields may be accomplished as single-point or multipoint grounding. Factors that influence the selection of single-point or multipoint grounding include the interference signal frequencies involved, the length of the transmission line, and the relative sensitivity of the circuit to high- or low-impedance fields.

(1) Single-point shield grounding. For multilead systems, each shield may be grounded at a different physical point as long as individual shields are insulated from each other. Single-point grounding is more effective than multipoint shield

grounding only for short shield lengths. Single-point grounding is ineffective in reducing magnetic or electrostatic coupling when conductor-length-to-wavelength ($L/\lambda$) ratios are greater than 0.15, where the wavelength is that of the highest frequency to be used (or the highest frequency interference to be expected) on the wire or in the system.

(2) Multipoint shield grounding. For $L/\lambda$ ratios greater than 0.15, multipoint grounding at intervals of $0.15\lambda$ is recommended, for the shield can act as an antenna that is relatively efficient at $\frac{L/\lambda}{4}$ when one end is grounded. When grounding the shield at intervals of $0.15\lambda$ is impracticable, shields should be grounded at each end. Multipoint shield grounding is effective in reducing all types of electrostatic coupling, but is subject to failure if large ground currents exist. In general, multipoint shield grounding solves most problems, but in audio circuits single-point shield grounding may be more effective because of the ground current problem.

c. Printed Circuit Boards. A requirement often arises that necessitates the grounding of hot shields on printed-circuit boards. For maximum shielding and isolation, shields on printed-circuit boards should be grounded directly to the main chassis, independent of any grounds located on the printed-circuit board. A typical problem would be the grounding of a shield for an rf transformer located on a printed-circuit board. The purpose of the shield is to prevent stray flux lines of the transformer from coupling into other circuits. Grounding the shield to the printed-circuit board permits the conduction of stray currents into the printed board circuitry, thus causing secondary interference problems. The correct way of grounding this type of shield is directly to the main chassis, as illustrated on figure 2-6. The grounding of the shield is completely independent of the rest of the circuitry.

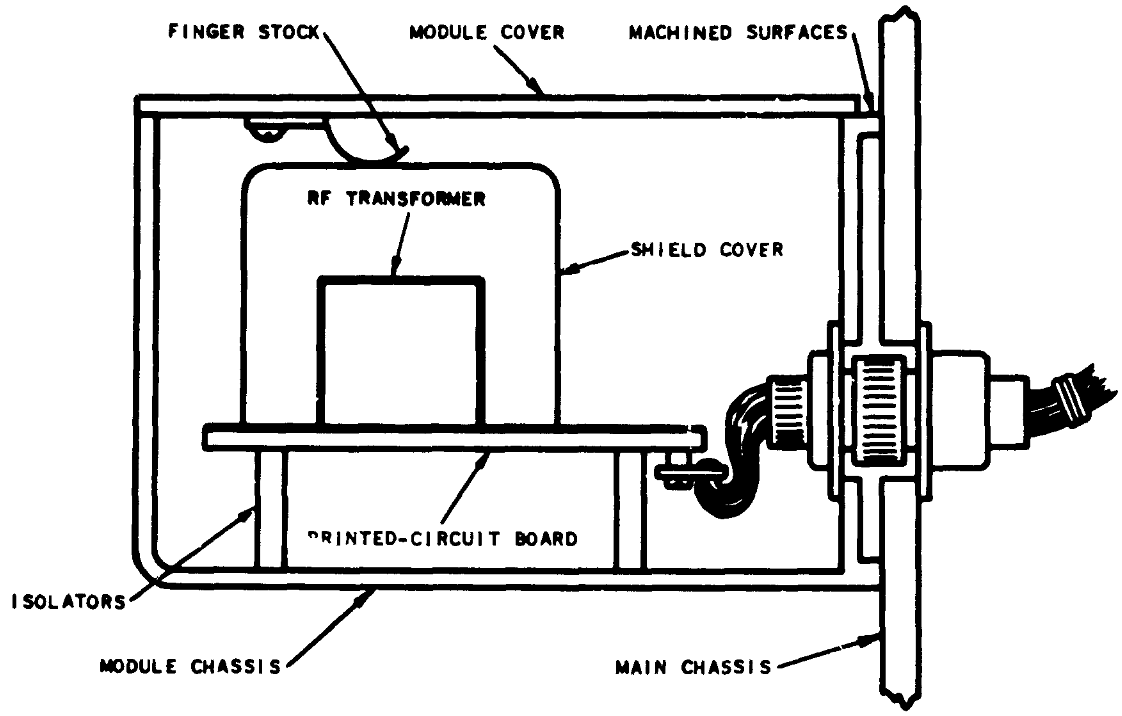



Figure 2-6. Direct Use of Chassis for Good Ground

## 2-6. Circuit Grounding

If the electronic ground plane in equipment exhibits low-impedance characteristics for all frequencies encountered in the system, it may be used as a common ground return, eliminating the need for individual wires. Under these conditions, the potential difference between any two points in the ground plane should not be of sufficient magnitude to cause generation and/or conduction of interfering signals to the input terminals of sensitive circuits. However, if the distances between two points are great, some potential difference will exist. These voltages must be considered when defining the permissible ambient noise level in the system and when determining the expected signal-to-noise ratio of the signal transmission circuits. Each electronic circuit contributes its own ground currents. Any ground return path that goes around corners or crosses other return paths may cause intercircuit or interstage coupling (fig. 2-7). Circuits using the ground plane for return currents may create a voltage that causes interference at the input terminals of sensitive circuits. The magnitude of this interference voltage is dependent upon impedance between the circuit ground points and the ground plane, and the current in the ground plane (fig. 2-8). The current (I) of the low-impedance circuit produces a potential ($IZ_{gp}$) in the high-impedance circuit. The simplest and most direct approach to this problem is to arrange circuit components physically so that ground return paths are short and direct and have the fewest possible crossings. In this way, the intercircuit coupling of these ground currents will be kept to a minimum. The effect of ground potential can be cancelled by electrically isolating the circuits. An example of this technique is the isolation transformer shown on figure 2-9. This method is especially effective at audio and low radio frequencies. Above these frequencies, its effectiveness progressively diminishes because, as the frequency of ground potential increases, more coupling paths appear. At radio frequencies, this potential can be considered as a voltage source in series with the signal return circuit and should be considered when designing critical low-level circuitry (fig. 2-10). A typical design for an electronically coupled circuit that considers the ground potential is shown on figure 2-11. Because of the plate characteristics of a pentode, a change of plate voltage has little effect on plate current. Thus, when the ground potential is coupled to the circuit through $R_K$, although it changes the plate voltage ($E_p$), it has very little effect on plate current ($i_p$).

## 2-7. Power Supplies

The power ground and signal ground should be isolated from each other throughout the chassis to minimize the possible coupling of signals from any one type of ground line to any other type of ground line. The application of the following techniques can often avoid potential problems:

1) Incorporate, where possible, individual ground paths for ac voltages, dc voltages, and signals

2) Connect a ground path to the largest conductor (lowest impedance) by as direct a route as possible

3) Utilize several arterial ground paths to the supply common point, as opposed to one superground bus

4) Avoid multiended ground buses or lateral ground loops

5) Have as few series connections (solder joints, connectors) as possible in a ground bus, and make sure that they are good, solid electrical connections

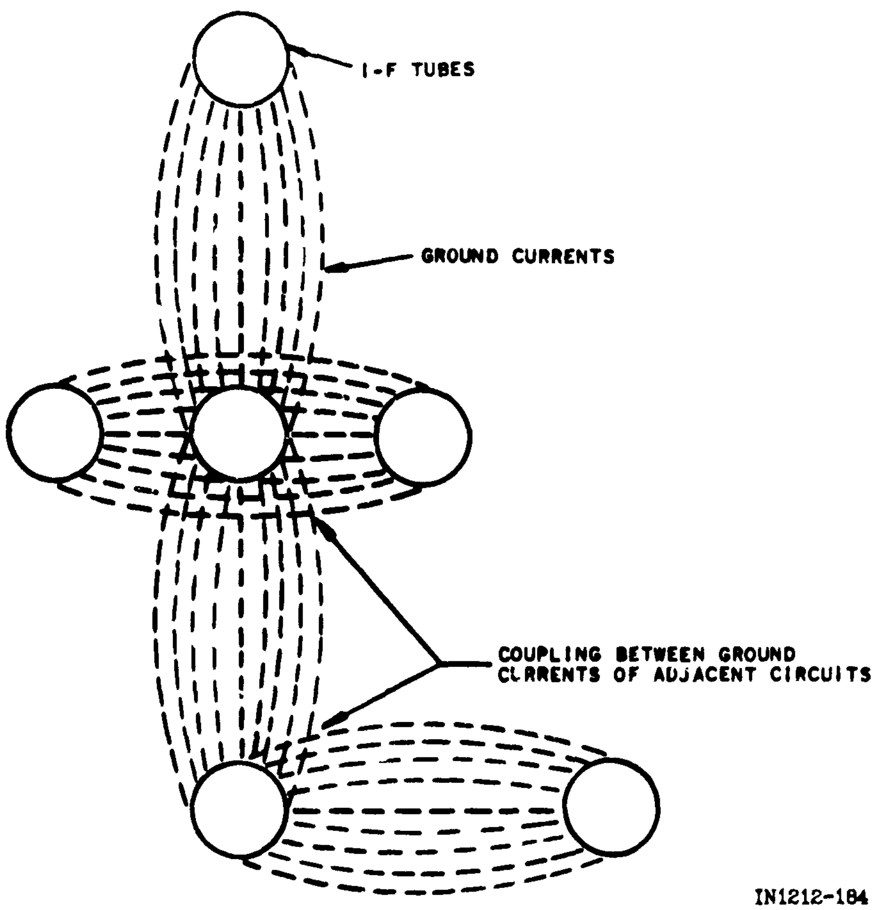



Figure 2-7. Example of Intercircuit Coupling

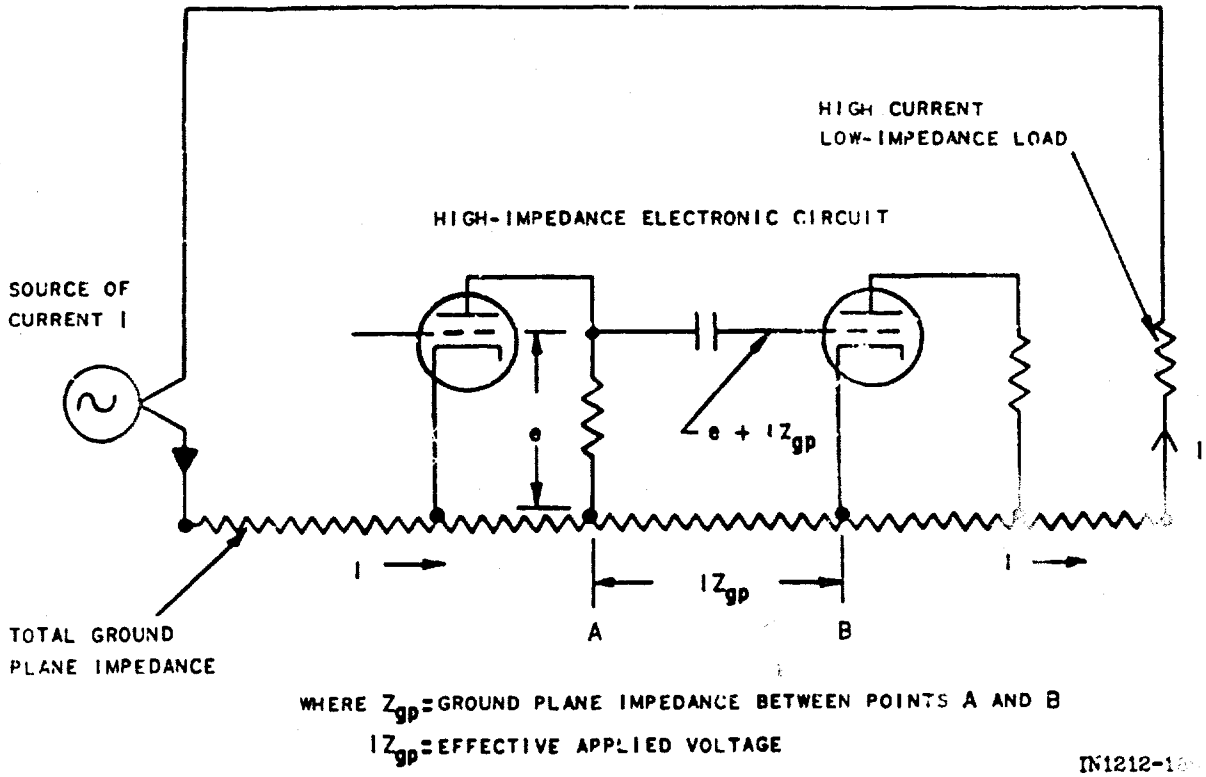



Figure 2-8. Effect of Circulating Currents on a Typical Circuit

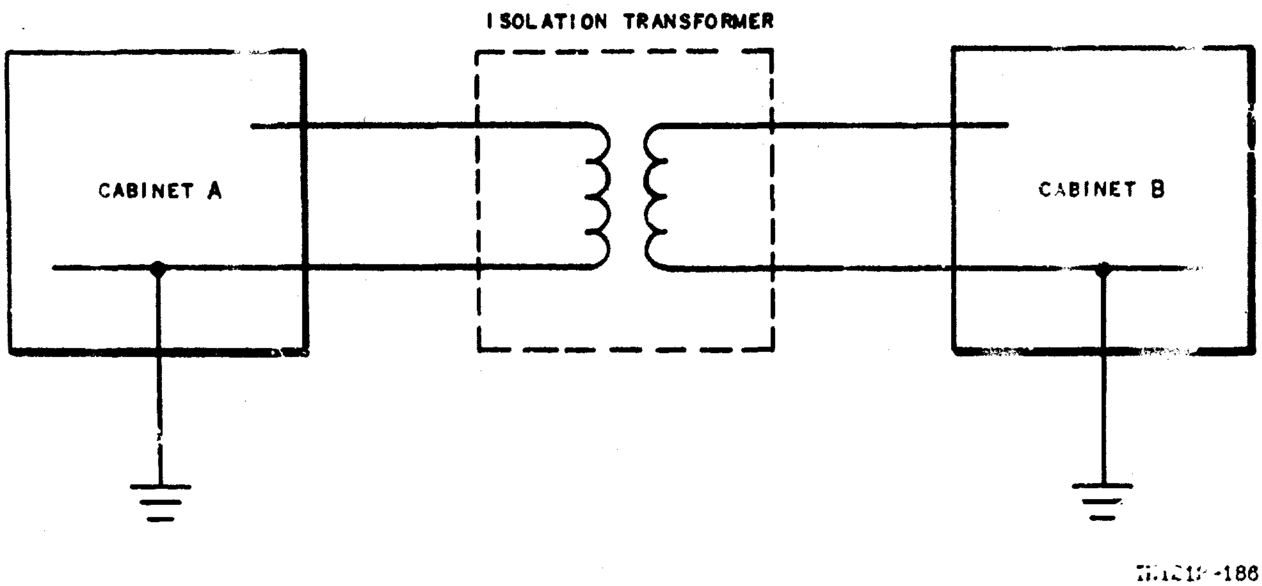



Figure 2-9. Isolation Transformer Technique for Minimizing Ground Potential

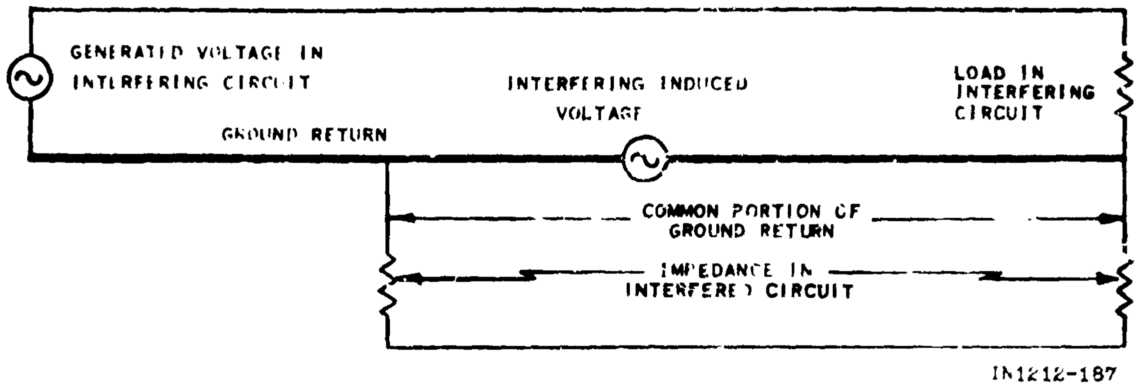



Figure 2-10. Equivalent Circuit of Ground Noise in
Electronically Coupled Circuits

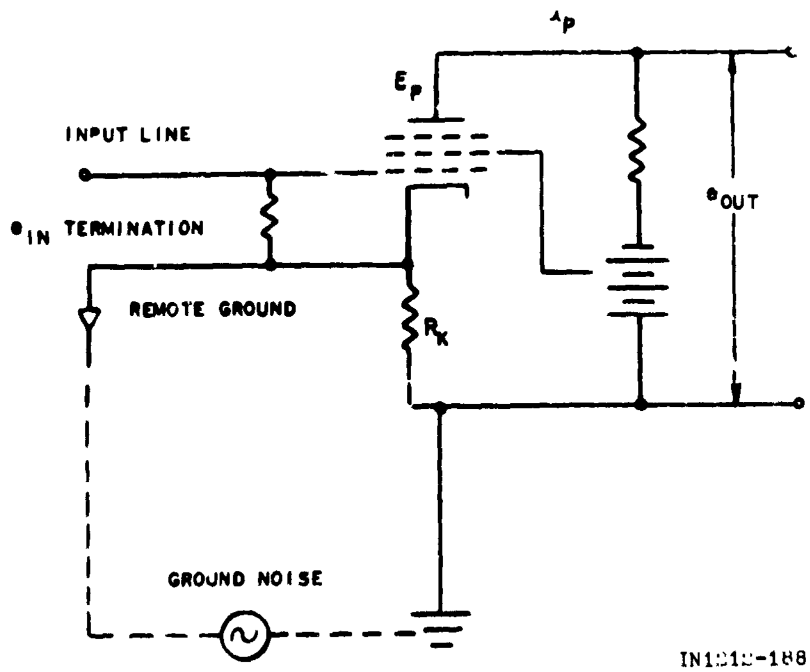



Figure 2-11. Electronically Coupled Circuit With
Ground Noise

## Section III. BONDING

### 2-8. General

Bonding is essential to the prevention, control, and/or elimination of interference. Inadequate bonding frequently contributes to poor equipment performance; improved bonding almost always results in a reduction of interference.

### 2-9. Theoretical Discussion

a. A bond is an electrical union between two metallic structures used to provide a low-impedance circuit between them. Bonding is the procedure by which the housing or structure of a subassembly or component is electrically connected to another structure, such as the frame of an electrical machine, or chassis of an electronic assembly. Because the reason for bonding two or more units together is to simulate electrically a single homogeneous structure and to prevent development of electrical potentials between individual metal structures, it is important that the bond present a low-impedance path to frequencies capable of causing interference.

b. The effectiveness of a bond at radio frequencies is neither fully dependent upon nor measurable only in terms of its dc electrical resistance; especially at high frequencies, where lengths of bonding jumpers tend to approach the wavelengths of undesirable electromagnetic radiation. When this occurs, and a bonding jumper presents a high-impedance path, there is a potential drop across the bond, and the metal structures connected by the bond remain at different potentials. As a result, the metal structures do not function effectively as shields and fail to limit interference radiation from and susceptibility to circuits within. As it is more convenient to measure the dc resistance rather than the ac impedance of a bond, dc measurement is often employed as an indication of low-frequency bonding effectiveness. This may be accomplished with a resistance bridge. At high frequencies, however, bond effectiveness is best determined by means of impedance measurements because bond capacitance and inductance become significant and may result in relatively high rf bond impedances, despite low dc resistance readings. The equivalent circuit of a bond strap and its impedance as a function of frequency are shown on figure 2-12. In practice, dc resistance measurements are utilized to detect grossly defective bonds, and to determine quickly, by comparison with manufacturer's test data, whether or not bonds on existing equipment have deteriorated in the field. The dc resistance of an adequate bond should be between 0.00025 and 0.0025 ohm. In addition to impairing shielding effectiveness, high-impedance bonding jumpers may radiate rf energy.

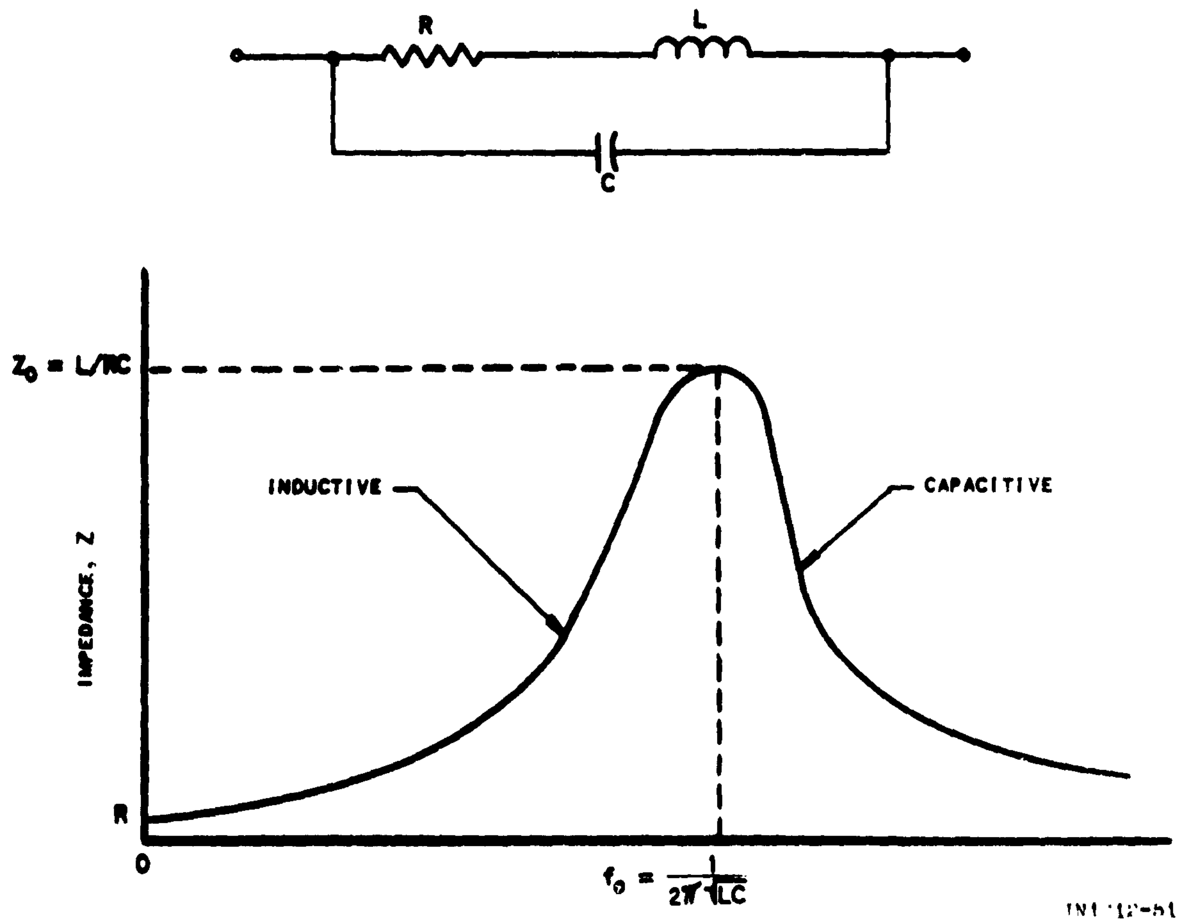



Figure 2-12. Bonding Strap Impedance Characteristics

c. In designing and establishing bonding criteria for specific applications, it is necessary to consider a variety of factors, such as interference frequency spectrum and maximum allowable bonding impedances for frequencies within a specific range. Of prime importance are such physical characteristics of the bonds selected as size, strength, fatigue resistance, corrosion resistance, resistivity, and temperature coefficients. It is the design engineer's responsibility to provide bonds that will not deteriorate appreciably even when equipment is subjected to adverse environmental conditions. Bonds may be affected by electrolytic action between the metals used and their surroundings. An excellent bond at time of fabrication may actually become a serious interference source shortly afterwards if proper precautions have not been taken.

d. Bonding jumpers should preferably be flat, solid straps to provide large surface areas for low rf impedance (rf currents flow along conductor surfaces). The measured rf impedance of a typical flat bond-strap at frequencies up to 30 mc increases almost linearly with frequency; such impedance is due almost entirely to the self-inductance of the strap. The capacitance between the bonded members is in parallel with the inductance of the bond strap; and the bond strap has the characteristics of a parallel capacitance-inductance circuit operating far below its resonant frequency. At the frequency of self-resonance, the rf impedance of such a parallel capacitance-inductance circuit is relatively high compared to the dc resistance, and effectiveness of the bond strap is at a minimum.

e. At high frequencies, where length of a bond strap is an appreciable part of a wavelength, the bond strap becomes equivalent to an rf transmission line; and impedance varies periodically from a minimum to a maximum with increasing frequency. When this happens, the bonding strap can act as a radiator of rf energy because it has rf current flowing through it and rf voltage across it. The sheet or metal part supposedly bonded to the main structure may be considered as the flat top of an antenna that is being fed by the bonding strap. These effects are reduced by keeping inductance as low as possible. Straps of minimum length and high width-to-thickness ratio should be used.

## 2-10. Types of Bonds

There are two classifications of bonds: direct and indirect. The most desirable of these is the direct bond. This term is applied to permanent, metal-to-metal joints such as are provided by welding or brazing. Indirect bonds, or flexible metal straps, are used when metals to be bonded cannot be placed in direct contact; for example, when there is a need for motion between bonded members.

a. Direct Bonds. Direct bonds include permanent metal-to-metal joints formed of machined metal surfaces; or with conductive gaskets held together by lock-threaded devices, riveted joints, tie rods, or pinned fittings driven tight and not subject to wear or vibration. The best bonded joint is formed by welding, brazing, or sweating. Soldering is not a good method of direct bonding because soldered joints have appreciable contact resistance. Basic requirements for direct bonding are that good metal-to-metal contact be provided for the life of the joint, and that precautions be taken to seal the joint against moisture that would cause galvanic corrosion. Dissimilar metals in direct contact should be avoided. Screw threads are never considered adequate bonding surfaces. In particular, sheet-metal type screws are inadequate for use in bonding. If two structural members are held together by screws, the impedance between them is usually comparatively high unless good direct contact is maintained.

b. Indirect Bonds. When a direct bond is not practical, the designer should select an indirect bond. A good indirect bond is one that presents a low impedance throughout the interference spectrum and retains its usefulness for an extended period of time. An indirect bond is usually a bond strap or jumper, mechanically held by means of bolts, rivets, welding, brazing, or sweating. Tooth-type lockwashers are used with bolt fasteners to ensure no deterioration of the metal-to-metal contact of bond-strap connections. The most significant feature of a bond-strap is its resiliency. When a solid strap is used, resiliency is determined by its material and thickness. Beryllium copper or phosphor bronze are often used and, under conditions of severe vibration, a corrugated strap often proves useful in preventing excessive damping and in achieving maximum service life. Figure 2-13 shows a typical bond strap bolted into position. Good metal-to-metal contact at the point of bonding is required for efficient operation, and any discussion of corrosion is not intended to compromise this requirement.

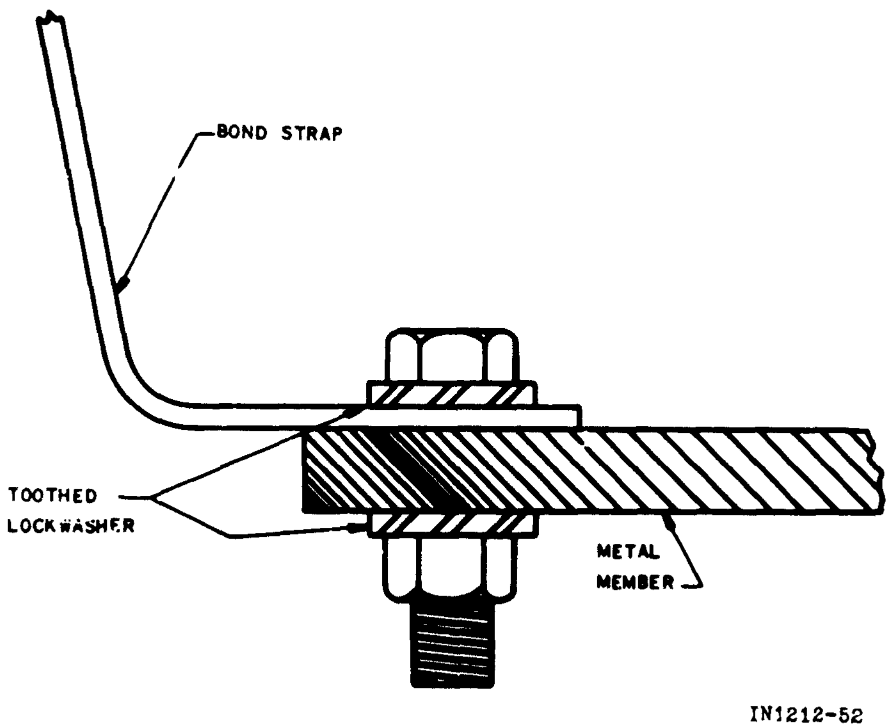



**Figure 2-13. Recommended Bond Strap Bolting Installation**

(1) Bonding jumpers. Bonding jumpers are short, round, braid conductors for application where the interference frequency is below a few megacycles. They are generally used in low-frequency devices and where the development of static charges must be prevented.

(2) Bond-straps. Bond-straps are either solid, flat, metallic conductors, or a woven braid configuration where many conductors are effectively in parallel. Solid-metal straps are generally preferred for the majority of applications.

Braided or stranded bond-straps are not generally recommended because of several undesirable characteristics. Oxides may form on each strand of non-protected wire and cause corrosion. Because such corrosion is not uniform, the cross-sectional area of each strand of wire will vary throughout its length. The nonuniform cross-sectional areas (and possible broken strands of wire) may lead to generation of interference signals within the cable or strap (fig. 2-14). Broken strands may act as efficient antennas at high frequencies, and interference may be generated by intermittent contact between strands. Solid straps are also preferable because of lower self-inductance. The direct influence of bond-strap construction on rf impedance is shown on the graph of figure 2-15, where the impedances of two bonding straps and of No. 12 wire are plotted against frequency. The relatively high impedance at high frequencies illustrates that there is no adequate substitute for direct metal-to-metal contact. A rule of thumb for achieving minimum bond strap inductance is that the length-to-width ratio of the strap should be 5:1 or less. This ratio determines the inductance, the major factor in the high-frequency impedance of the strap.

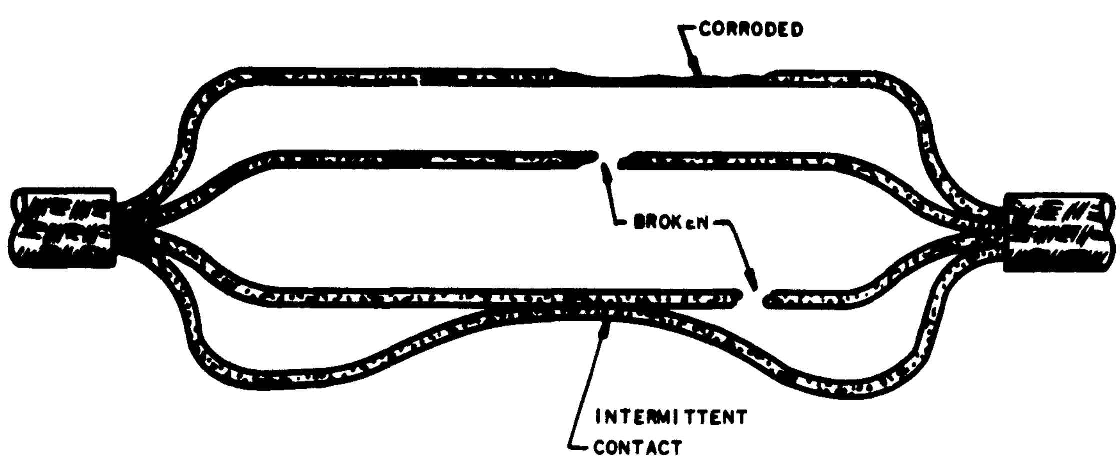



Figure 2-14. Exploded View, Individual Strands of Braided Cable

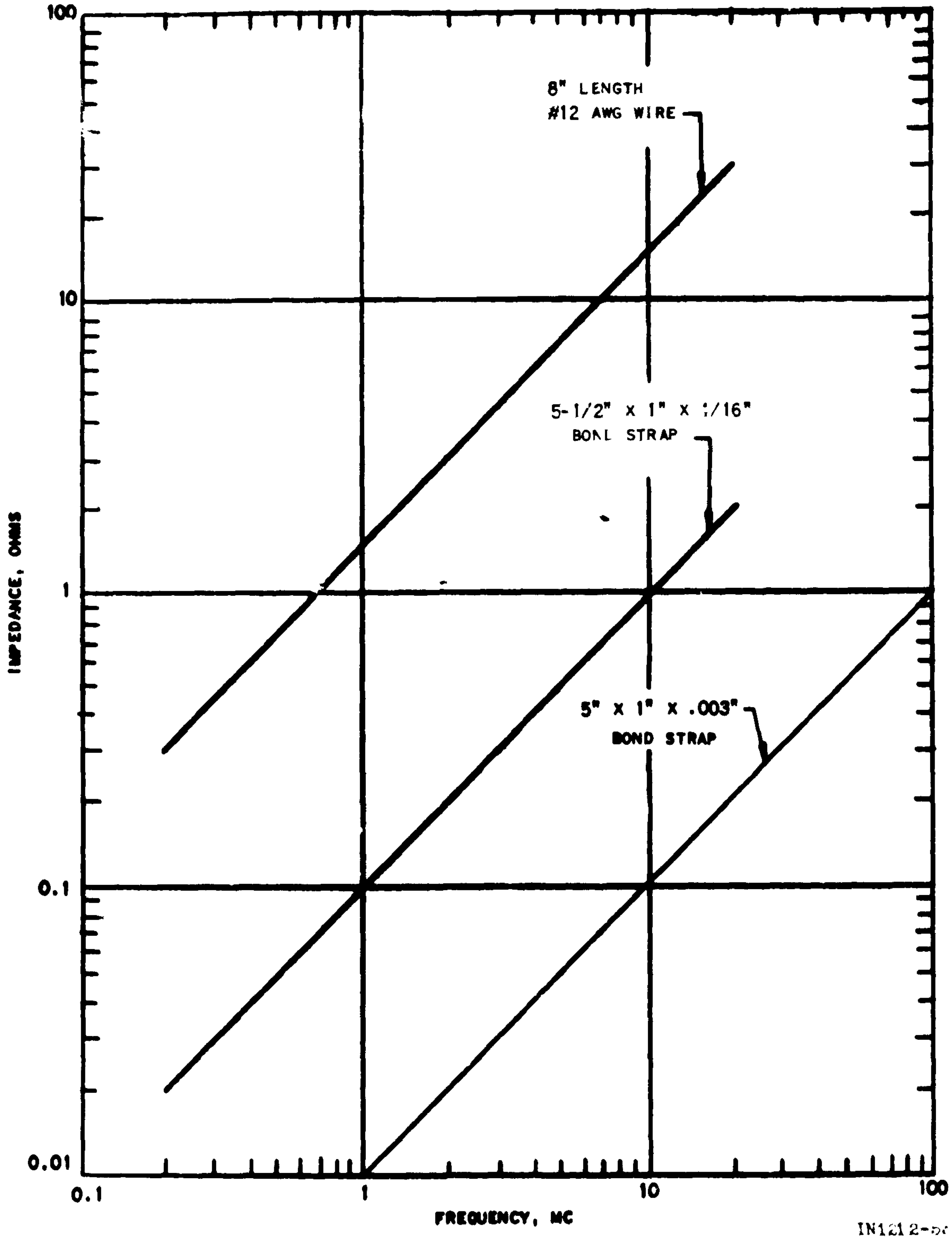



Figure 2-15. Impedances of Bond-Straps and No. 12 AWG Wire

## 2-11. Bonding Metal Selection and Bond-Strap Finishes

a. The choice of material for a given bonding application is usually dictated by consideration of the metals being bonded and the environment within which the bond must exist. In bonding, the necessity for joining dissimilar metals is frequently unavoidable. In such cases, corrosion becomes an important consideration. Factors contributing to corrosion are the relative closeness of metals in the electromotive series and the amount of moisture present. Corrosion is attributed to two basic electrochemical processes: galvanic and electrolytic corrosion.

b. Several methods can be employed for minimizing or preventing corrosion and its adverse effects on bonding. One is to use metals low on the activity table, such as copper, lead or tin (table 2-1). Where members of the electrolytic couple are widely separated on the activity table, it is sometimes practical to use a plating such as cadmium or zinc, which helps to reconcile the dissimilarity. Thin, bimetallic plates, formed by mechanical bonding of dissimilar metals cold flowed together under high pressures, have been used to interconnect two structural units of dissimilar metals. Where bimetallic plates are to be used, the junctures of the two metals are normally covered with a protective coating, such as grease or polysulphate, to exclude moisture and retard corrosion. This coating reduces the area of metal exposed to an electrolyte, thus reducing corrosion. If bonding is such that corrosion is likely to occur, the bond should be designed as a replaceable element, such as a jumper, plate, separator, or washer.

c. Acceptable contact surface materials that may be used to fasten bonding jumpers to structures are indicated in table 2-2. The arrangement of the metals listed in this table is in the order of their decreasing galvanic activity when exposed to an electrolyte. The screws, nuts, and washers to be used in making the connections are indicated as Type I, cadmium or zinc plated, or aluminum, and Type II, passivated stainless steel. Where neither type of securing hardware is indicated, Type II is preferred from a corrosion standpoint.

TABLE 2-1. ELECTROMOTIVE FORCE SERIES OF COMMONLY USED METALS[a]

| Metal | Electrode Potential (Volts) |
|---|---|
| Magnesium | +2.40 |
| Aluminum | +1.70 |
| Zinc | +0.762 |
| Chromium | +0.557 |
| Iron | +0.441 |
| Cadmium | +0.401 |
| Nickel | +0.231 |
| Tin | +0.136 |
| Lead | +0.122 |
| Copper | -0.344 |
| Silver | -0.798 |
| Platinum | -0.863 |
| Gold | -1.50 |

a - Select dissimilar metals so that if corrosion occurs, it will be in the replaceable components, such as grounding jumpers, washers, bolts or clamps, rather than structural members or equipment enclosures. When two different metals are in contact, the one higher in the electromotive-force series will be more affected by corrosion than the other. The smaller mass (generally the more easily replaceable) should therefore be made of the higher metal; for example, cadmium-plated washers are recommended for use with steel surfaces.

TABLE 2-2. METAL CONNECTIONS

| Metal Structure (Outer Finish Metal) | Connection for Aluminum Jumper | Screw Type[a] | Connection for Tinned Copper Jumper | Screw Type[a] |
|---|---|---|---|---|
| Magnesium and magnesium alloys | Direct or magnesium washer | Type I | Aluminum or magnesium washer | Type I |
| Zinc, cadmium, aluminum and aluminum alloys | Direct | Type I | Aluminum washer | Type I |
| Steel (except stainless steel) | Direct | Type I | Direct | Type I |
| Tin, lead, and tin-lead solders | Direct | Type I | Direct | Type I or II |
| Copper and copper alloys | Tinned or cadmium-plated washer | Type I or II | Direct | Type I or II |
| Nickel and nickel alloys | Tinned or cadmium-plated washer | Type I or II | Direct | Type I or II |
| Stainless steel | Tinned or cadmium-plated washer | Type I or II | Direct | Type I or II |
| Silver, gold, and precious metals | Tinned or cadmium-plated washer | Type I or II | Direct | Type I or II |

a - Type I is cadmium- or zinc-plated, or aluminum; Type II is stainless steel. Where either type is indicated as acceptable, Type II is preferred from a corrosion standpoint.

d. The possibility of galvanic and/or electrolytic action necessitates extreme care in assembling joints that serve as bonds. Surfaces should be absolutely dry before mating, and should be held together under high pressure to minimize the possibility of moisture entering joints. The use of number 7/0 garnet finishing paper or equivalent is recommended to remove paints, anodic films, and oxides from surfaces. Care must be taken not to remove excessive metal under the protective finish. Abrasives, such as emery cloth or sandpaper, cause corrosive action because their particles embed themselves in the metal; they therefore should not be used. The contact area should be brushed clean; it should be about 1-1/2 times greater than the area necessary for actual mounting. After a joint (free of moisture) is assembled, the periphery of the exposed edge should be sealed with suitable grease or a polysulphate coating.

2-12. Bonding Applications

a. Shock Mounts. A frequent application for which indirect bonding is the only suitable type is that involving shock-mounted equipment. The designer should consider the degree of relative motion to be expected between two surfaces to be bonded, the characteristics of the materials involved, and the frequency range over which the bonding is expected to be effective. A typical shock mount is shown on figure 2-16. The application of a bond-strap to a vehicle engine is shown on figures 2-17 and 2-18. The resiliency of the bonded mount should be determined by characteristics of the mount, not of the bond-strap. The strap should not significantly dampen the shock mount, and where necessary, it should be corrugated to withstand severe and continued vibration. Where interference suppression is desired in the vhf range and higher, two bond straps across each shock mount should be used. This arrangement reduces the inductance of the bond to half of its former inductance and increases the resonant frequency of the strap. The use of tooth-type lock washers is preferable so that perforation of any nonconductive coating (with improved electrical contact) is assured. Where severe environments are involved, joints should be protected after tightening with a suitable grease or polysulphate coating to preclude corrosion at contact surfaces.

b. Rotating Joints. It is often necessary to bond shafts of rotating machinery to prevent accumulation of static charges. Bonding is generally accomplished by use of a slip ring and brush assembly, or a phosphor-bronze finger riding directly on the shaft. (See Section X of Chapter 3).

c. Tubing Conduit. The outer surfaces of long spans of conduit or shielded cable may be high-impedance paths for interference currents from external sources. To minimize this possibility, such spans should be properly bonded to structures at both ends and at several intermediate points. Ordinary clamps cannot be used to bond flexible conduit since the required pressure on a comparatively small surface area of the conduit may be sufficiently high to compress or collapse it. To overcome this, a flared split-sleeve is fitted around the flexible conduit. This sleeve distributes the high pressure of the bonding clamp over a large area, thereby exerting low pressure on the conduit (fig. 2-19A). Contact is further improved by soldering the sleeve to the conduit, material permitting through several holes in the sleeve provided for this purpose. Figure 2-19B illustrates a method for bonding rigid conduit to a structure through supporting attachments. The conduit or tubing, to which bonding clamps are attached, should be cleansed of paint and foreign material over the entire area covered by the clamps. All insulating finishes should be removed from the contact area before assembly, and anodized screws, nuts, and washers should not be used to attach contacting parts. If, in bolting the clamp to the bonding surface, a tooth-type washer is used, protective coatings, unless very thick or tough, need not be removed from the surface because the points of the washer will penetrate to the bare metal.

d. Hinges. Hinges do not provide a path for electrical conductivity or an rf shield. Where hinges must be used, it is necessary to accomplish bonding by other means. Figure 2-20 shows a typical configuration for bonding hinges. Flexible bonding-straps, made of thin metal, are separated along the hinges by no more than 2 inches.

e. Cable Trays. Cable trays should be utilized as part of the overall system bonding scheme. Each section of each tray should be bonded to the following section to provide a continuous path (fig. 2-21). The trays should also be connected to equipment housings by wide, flexible, solid bond-straps. Such connections reduce the level of interference propagated into the equipment, thus precluding any difference of potential between equipment housings. A typical example of cable tray bonding is shown on figure 2-22.

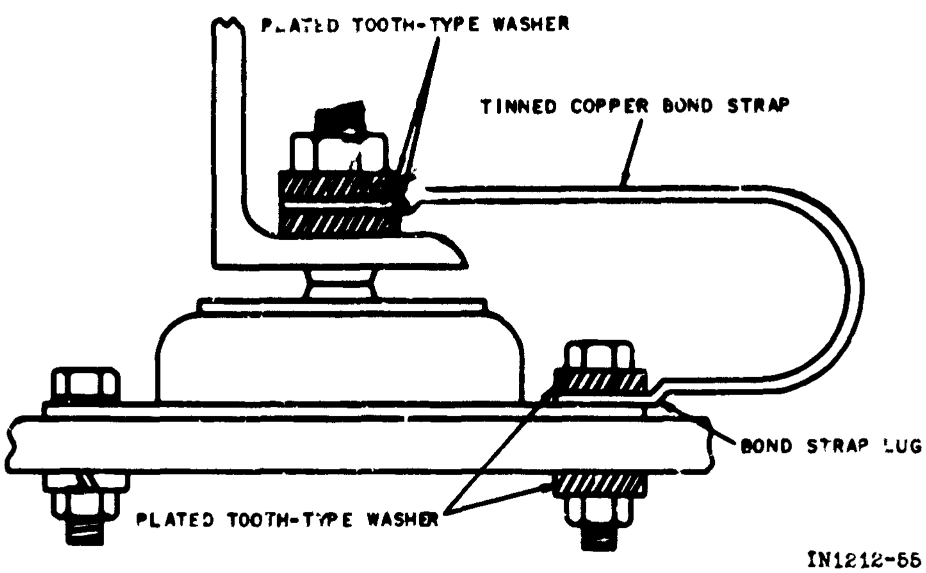



Figure 2-16. Typical Shock Mount Bond

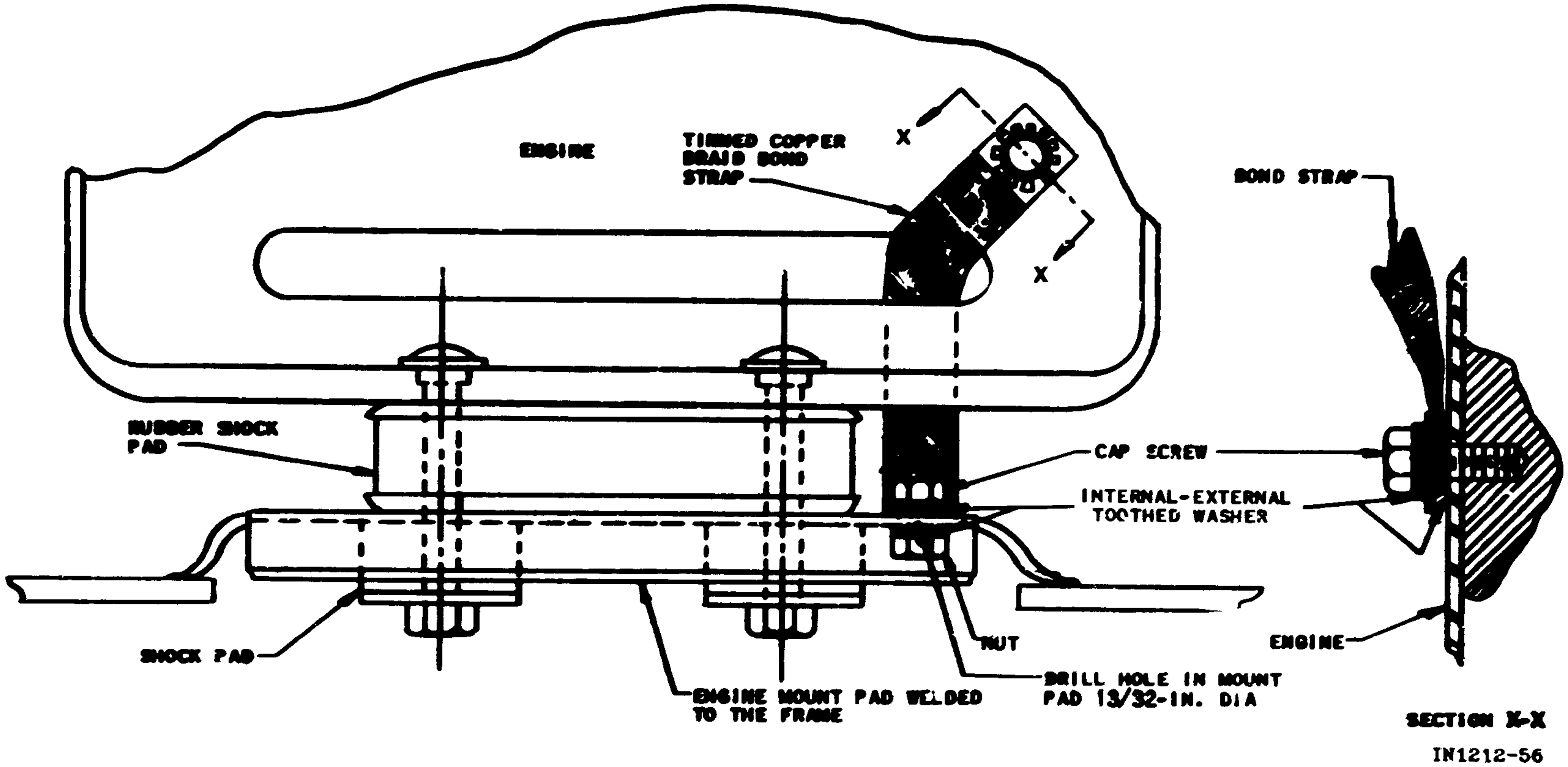



Figure 2-17. Bonded Engine Shock Mount - Front

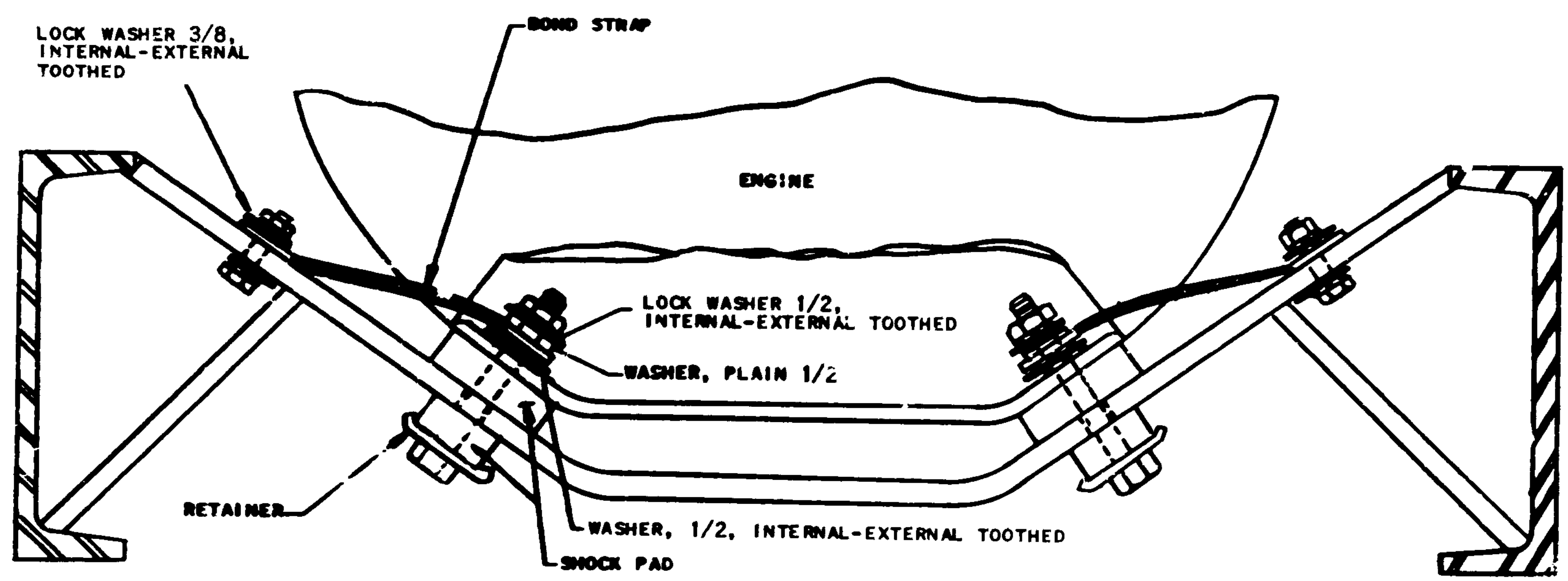



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Figure 2-18. Bonded Engine Shock Mount - Rear

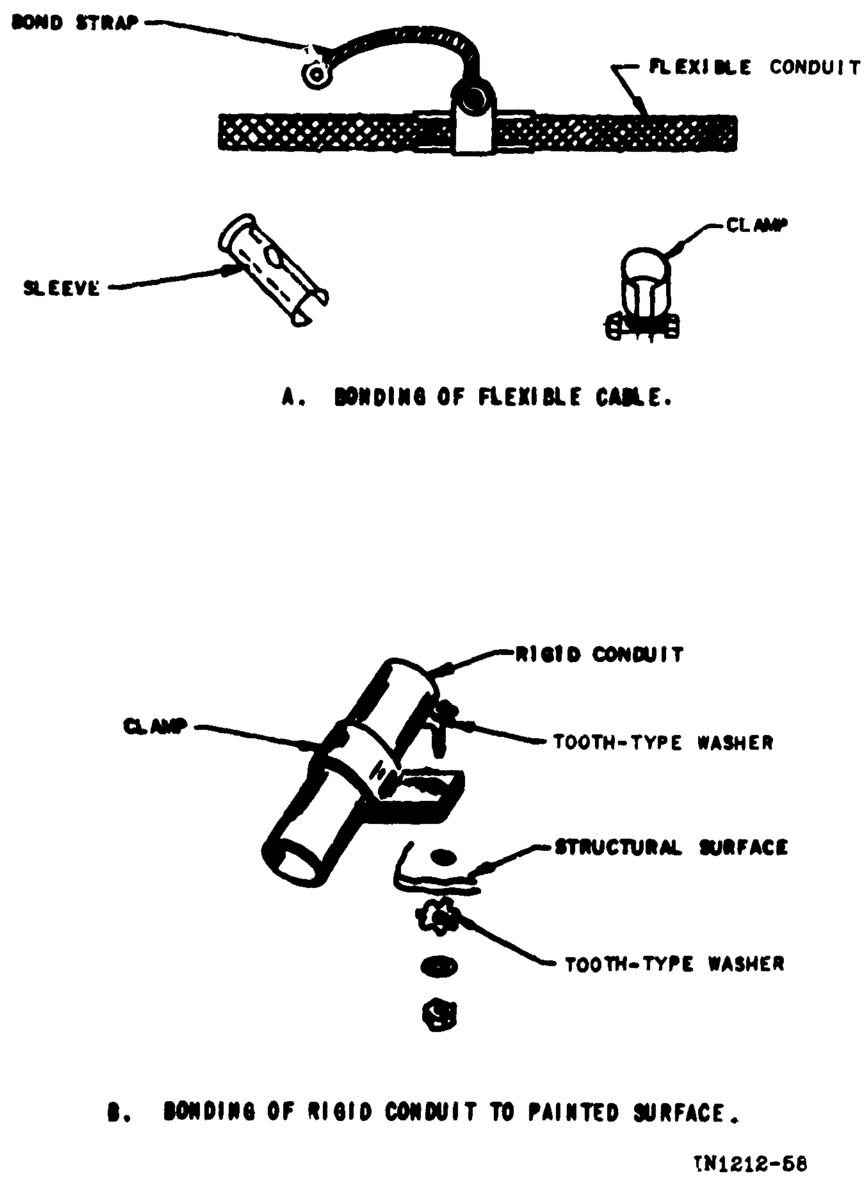



Figure 2-19. Cable and Conduit Bonding

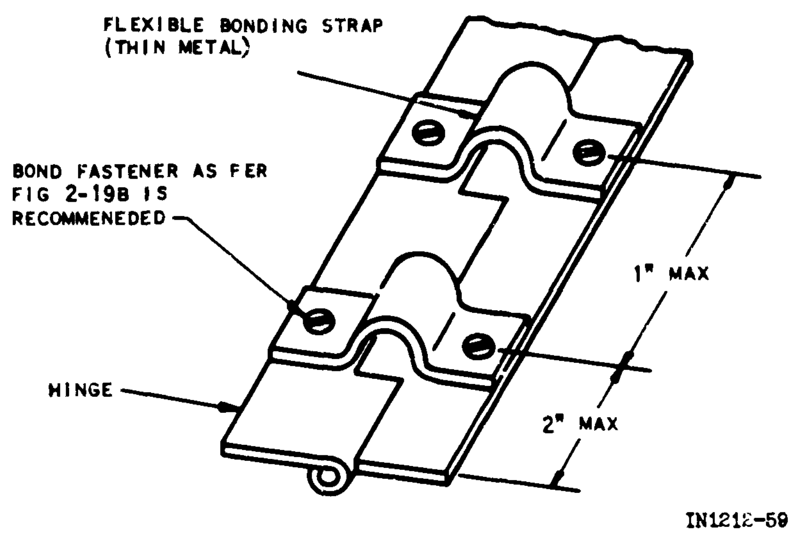



Figure 2-20. Bonding of Hinges

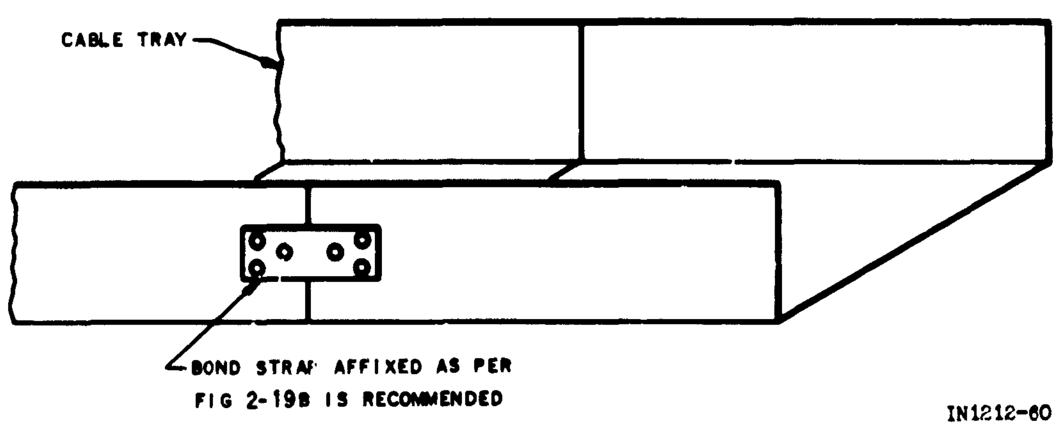



Figure 2-21. Cable Tray Section Bonding

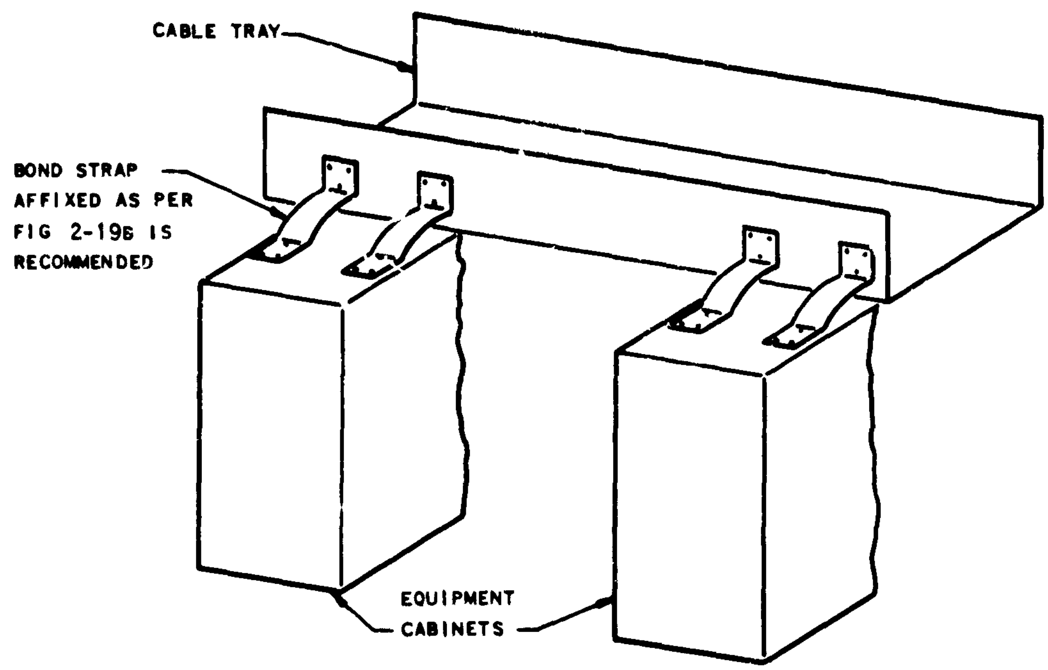



Figure 2-22. Equipment Cabinets Bonded to Cable Tray

f. Bonding Practices Summarized.

1) Permanent-type bonds are more reliable than the semipermanent type, and are therefore preferred

2) Direct-type bonds, such as formed by individual welded, sweated, or brazed joints are, in general, bonds of lower impedance than indirect types, and are therefore preferred

3) Where bond joints exist between dissimilar metals, finished bond joints should have a protective coating such as a suitable grease or polysulphate to exclude moisture and retard corrosion

4) Bonds should afford good metal-to-metal contact over the entire mating surfaces of the bond joint. The mating surfaces should be clean and free from any nonconductive finishes. Bare, clean, metal-to-metal contact will ensure a low-impedance connection between mating surfaces

5) Indirect bonding conductors should preferably be in strap form, broad in width, thin, and as short in length as possible to afford desirable low-impedance electrical connections at radio frequencies. The length-to-width ratio of the bond straps should be less than 5:1

6) The strap type of bond connection provides flexibility, sometimes necessary because of vibration, expansion, contraction, hinges, and equipment misalignment arising from normal fabrication and installation tolerances

7) Where bonds must afford shielding integrity, permanent-type metal-to-metal joints, afforded by welding, brazing, or sweating, are preferred to semipermanent joints that depend on clamping pressures and/or conducting gaskets

8) A soldered bond-joint should not depend on the solder for mechanical strength. The parent mating materials of the bond should be mechanically jointed by other means such as bolting or riveting

9) Bonding connections should be located in protected and accessible areas, where practical, to permit ready inspection and replacement if necessary

## Section IV. SHIELDING

### 2-13. General

Shielding may be required in electrical and electronic equipment to prevent the equipment from propagating interference and to protect the equipment from the effects of interference propagated by other electronic devices. In the design of electronic equipment shielding, the mechanical and electrical design engineer should consider the following:

1) Interference specification requirements

2) Expected potential radiation emanating from the subassemblies, assemblies, and complete units

3) Types of energy fields involved and the most satisfactory and economical materials to suppress or eliminate their effects

4) Requirements for corrosion prevention, continuity of surface contact, and heat dissipation

A shield should be of solid metal covering, flexible metal conduit, or mesh screen. Such electromagnetic shielding can be used in a variety of ways and forms, ranging from small modules to enclosures for high-power equipment. One of the most difficult problems occurs when many transmitters, receivers, and/or other sensitive equipment are installed close together. On vehicular carriers, the problem is intensified because increased electronic demands require integration of more electronic functions within one compact enclosure. The effectiveness of structural shielding can range from 20 to 100 db. This shielding is generally not sufficient to protect receiving antennas from undesirable signals generated within the enclosure. An additional 30 db of attenuation can be obtained by shielding the antenna wire, or by using coaxial cable or waveguide.

### 2-14. Shielding Effectiveness

a. Shielding effectiveness is a measure of the total ability of a material to prevent propagation of electromagnetic energy. In the classical shielding theory, which compares a shield to a transmission line, the radiated field is considered to be reflected and attenuated in passing through metal (fig. 2-23A). This theory is analogous to the theory of propagation of traveling waves on a transmission line. It is possible (following the transmission line analogy) to compute the attenuation through the shield and the reflections that will be present at each surface. Empirical data, gathered over a number of years, has shown that the classical shielding theory accurately describes the shielding process. The shielding effectiveness is represented schematically on figure 2-23B. The mathematical relationship is:

$$SE = R + A + B \tag{2-1}$$

where: $SE$ = total shielding effectiveness (db)

$R = R_1 + R_2$ = reflection radiated power loss of the first and and second boundary (db)

$A$ = absorption power loss (db)

$B$ = B-factor (db), which is neglected if A is greater than 10 db

$P_1$ = incident radiated power (dbw)

$P_2$ = exiting radiated power (dbw)

Expressing the law of conservation of energy:

$$P_2 = P_1 - R_1 - A - R_2 - B$$

$$P_1 - P_2 = R_1 + R_2 + A + B$$

$$P_1 - P_2 = R + A + B$$

$$SE = P_1 - P_2 = 10 \log \frac{P_1}{P_2}$$

For convenience, the above equations are expressed in terms of incident and emitted power. However, at low frequencies, when operating in the induction field, it is more convenient to work with electric or magnetic field intensities. In this case:

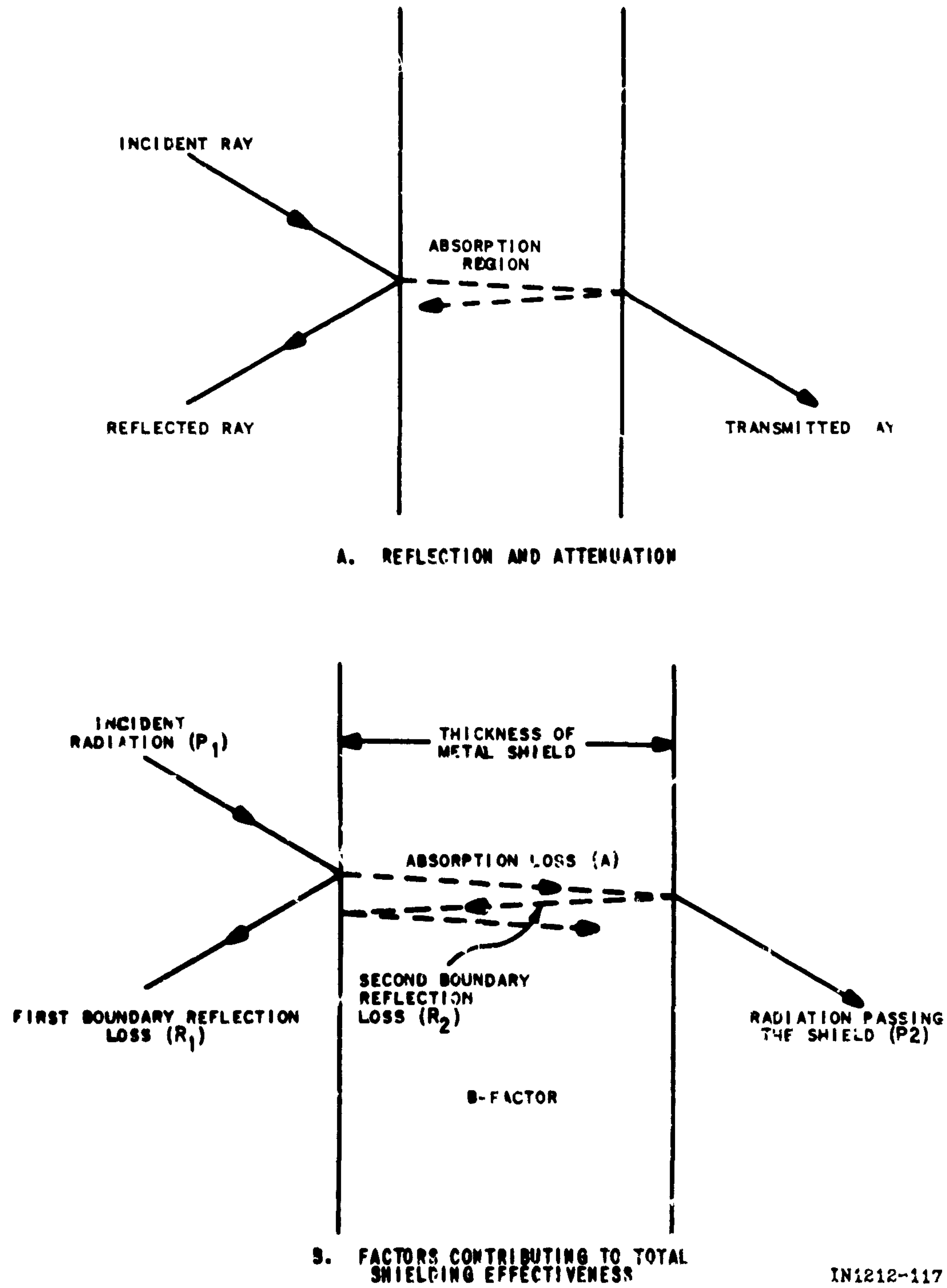



Figure 2-23. Metal Shielding Effectiveness

$$SE = 20 \log \frac{E_1}{E_2} = 20 \log \frac{H_1}{H_2}$$

These losses are a function of frequency, thickness of material, resistivity, permeability, and conductivity. The reflection losses vary with the characteristic wave impedance of the electromagnetic field and may also vary with distance. Magnetic fields occur in the vicinity of coils or small loop antennas. Because reflection losses for magnetic fields are small for most materials, magnetic shielding depends primarily on absorption losses. Electric fields are readily stopped by metal shields because large reflection losses are easily obtained. The absorption loss, which is essentially independent of the wave impedance, is the same for both electric and magnetic fields.

b. The total reduction in field intensity is caused by reflection and absorption losses. The shielding effectiveness of an enclosure is the sum of the reflection and absorption losses. Absorption loss represents the reduction in signal due to dissipation as it proceeds through the body of a shield; it is independent of the type of radiator emitting the signal. Reflection losses take place at the surfaces of a shield and vary with the ratio of the wave impedance emitted by the radiator to the intrinsic impedance of the shield. If the absorption loss is less than 10 db, then the B-factor must be calculated and added to equation 2-1. The B-factor can be neglected when the thickness of metal is sufficient to make the penetration loss more than 10 db. Shields that contain openings, such as copper screening or incompletely welded metal seams, introduce other complexities in that signals leak through the openings in addition to penetrating the shield material itself. Absorption losses can be very great for openings whose size is small compared to the wavelength of the energy passing through. They are similar to losses obtained with a waveguide below cutoff.

c. Tests made with small antennas (such as a 3-inch loop) measure shield quality only in the vicinity of the antenna location and usually give the shielding quality of a particular flaw. Measurements made with large antennas test the quality of a large area of the shield and integrate leakage obtained through that area.

2-15. Case and Front Panel Design

a. The metal walls of a vehicle aid in reducing the internal ambient level of interference signals from external sources and act as an interference barrier between internal and external equipment. The shielding effectiveness of the vehicle wall can range from 20 to 100 db. If this is not sufficient to protect the receiving antennas from undesired signals generated within the vehicle, additional shielding should be placed around the interference sources. Shields of up to 100 db effectiveness in reducing harmonic leakage from transmitter enclosures are obtainable.

b. Because it is difficult to shield against transmitter spurious radiation and receiver local oscillator radiation, adjacent pieces of equipment should be well shielded even if they do not generate high levels of undesirable energy. In such cases, it is generally most practical to provide equal shielding for each piece of equipment. To provide satisfactory shielding, an enclosure should have a shielding effectiveness of 50 to 100 db, depending upon intensity of undesired signals and whether an electric or a magnetic interference field is present. If all leads leaving an enclosure are well filtered, and the interference is of sufficiently high frequency to make the shield appear electrically thick, then for most purposes the interference is completely contained (or excluded) without regard to grounding. Grounding and physical wiring geography become important at lower frequencies where magnetic fields can readily penetrate shields. In general, the design objective should be maximum shielding effectiveness within limitations of weight, size, mechanical rigidity, and cost. An effective shield design prevents most interference energy from entering or leaving the susceptible space. Such a design uses a suitable barrier or metal enclosure, of proper thickness, with all discontinuities such as holes, seams, and joints sealed. Although good shielding materials are available, the greatest interference problem arises from the passage of interference signals through shielding discontinuities.

c. The effectiveness of shielding materials can be indicated in terms of their insertion loss expressed in decibels. Insertion loss is equal to shielding effectiveness when the shield is thin compared to the distance between test antennas. To determine the insertion loss of a given material, a shield is constructed of that material. Then, a signal generator is used to produce a test wave of constant frequency and amplitude. Readings are taken with a field intensity meter. Shielding effectiveness is calculated from the expression:

$$SE = 20 \log_{10} V_1 / V_2 \quad \text{(db)} \qquad (2\text{-}2)$$

where $V_1$ is the field intensity meter reading obtained with the radiating and receiving antennas located in free space insofar as possible and separated a specified distance. $V_2$ is the reading obtained with the antennas separated the same distance and with the shield inserted between them. The value for shielding effectiveness, derived from theoretical considerations, is usually much greater than that actually obtained by physical measurement. This fact is especially true at high frequencies (above 100 megacycles) where the plane wave field exists. The reason is that theoretical analysis is ordinarily based on the simple physical situation of a surface constituting a mathematical continuum -- that is, a smooth unbroken surface such as a sphere, a rectangular box, or an infinite flat plane. In fact, however, the practical surface is of finite dimensions and almost certainly has discontinuities. These discontinuities can be treated so that they do not entirely negate the shielding effectiveness, but some decrease from theoretically obtained SE values for continuous metal sheet should nevertheless be expected. At very low frequencies, the shielding effectiveness of these discontinuities usually is low due to greater depth of penetration. It improves as the frequency increases, and then it deteriorates again at very high frequencies where dimensions of openings in the shield become comparable to a half wavelength.

## 2-16. Shielding Design Fundamentals

The equations in this section are arranged and presented in a condensed form as an aid in the design of shielded enclosures. The terms used are defined as follows:

1) SE = Shielding effectiveness, representing the reduction of the level of incident electromagnetic energy (expressed in db) through the metallic shield in its path. Measurements are made in power, voltage, or current ratios

2) R = Total reflection loss, in db, from both surfaces, neglecting the effect of multiple reflections inside the shield

3) A = Absorption loss, in db, inside the shield

4) B = A positive or negative factor that need not be taken into account when A is more than 10 db. It is caused by the reflecting waves inside the barrier and is calculated in db. When a metallic barrier has an A of less than 10 db, it is designated as electrically thin. The use of electrically thin barriers should be avoided by using a different type of metal or a thicker specimen of the same metal

5) SE = R + A + B, when $A < 10$ db

6) SE = R + A, when $A > 10$ db

7) $Z_s$ = Intrinsic impedance of the metal

8) $Z_w$ = Wave impedance of incident wave in space

9) $\mu$ = Relative magnetic permeability referred to free space: 1 for copper, 1 for ferrous metals at microwave frequencies, and 200 to 1000 for ferrous metals at low frequencies

10) $\mu_o$ = Permeability of free space = $1.26 \times 10^{-6}$ henrys/meter, which is approximately $120\pi/v$

11) $\varepsilon$ = Permittivity of free space = $8.85 \times 10^{-12}$ farads/meter, which is approximately $\frac{1}{120\pi v}$

12) V = Velocity of light in free space = $3 \times 10^{8}$ meters/second = $f\lambda$

13) G = Relative conductivity referred to copper = 1 for copper, = 0.61 for aluminum, = 0.17 for iron

14) f = Frequency in cycles/second

15) $\lambda$ = Wavelength in meters/cycle

16) $\beta$ = $2\pi/\lambda$

17) $\omega$ = $2\pi f$

18) r = Distance from source to shield in meters

19) $r_1$ = Distance from source to shield in inches

20) t = Thickness of shield in mils

21) T = Thickness of shield in meters

22) E = Electric field component, or electric intensity in volts/meter

23) H = Magnetic field component, or magnetic intensity in amperes/meter

24) $\alpha$ = Attenuation constant of metal in nepers/meter

25) $\sqrt{\frac{\mu_o}{\varepsilon_o}}$ = Impedance of plane waves in free space = 377.6 ohms, = approximately $120\pi$

a. General Calculations.

(1) To calculate R:

$$R = 20 \log_{10} \left|(Z_s + Z_w)^2 / 4 Z_s Z_w\right| \text{ db}$$

$$Z_s = (1 + j) \sqrt{\mu f/2G} \times 3.69 \times 10^{-7} \text{ ohms} \qquad (2\text{-}3)$$

$$|Z_s| = \sqrt{\mu f/G} \times 3.69 \times 10^{-7} \text{ ohms}$$

R may be zero, positive, or negative, depending upon whether the ratio given is equal to, greater than, or smaller than unity, respectively. In all cases above 1 kc, R is positive. The corrected total reflection = R + B (algebraic sum); and may be zero, positive, or negative. In all cases above 1 kc, it is positive. B may be positive, negative, or it may equal zero. In most cases above 1 kc, it is negative. SE is positive and always greater than zero.

(2) To calculate $Z_w$ for high impedance electric fields, consider a very short nonresonant dipole where the length $<<\lambda$:

$$Z_w = \frac{E}{H} = \frac{1}{v\varepsilon} \times \frac{1 + j\beta r - \beta^2 r^2}{j\beta r - \beta^2 r^2}$$

when $r >> \lambda$, then:

$$Z_w = 1/v\varepsilon = 376.7 \text{ ohms} \qquad (2\text{-}4)$$

when $r << \lambda$, then:

$$Z_w = -j/\omega\varepsilon r \text{ ohms}$$
$$= -j\ 0.71 \times 10^{12}/fr_1 \text{ ohms}$$

For practical use, this calculation is a good estimate of the wave impedance when a rod or dipole of finite length (L) is used, providing that:

$$L << \mu << \lambda$$

(3) To calculate $Z_w$ for low impedance magnetic fields, consider a very small loop where the diameter $<<\lambda$:

$$Z_w = -\frac{E}{H} = V\mu_o \times \frac{j\beta r - \beta^2 r^2}{1 + j\beta r - \beta^2 r^2}$$

when $r >> \lambda$, then:

$$Z_w = V\mu_o = 376.7 \text{ ohms} \qquad (2\text{-}5)$$

when $r \ll \lambda$, then:

$$Z_w = + j\omega\mu_o r \text{ ohms}$$

$$= + j\ 0.2 \times 10^{-6}\ fr_1 \text{ ohms}$$

This calculation is accurate for finite loop diameters (D) when:

$$D \ll \mu \ll \lambda$$

(4) To calculate R for plane waves where $r > \lambda$, substitute in the previously given formulas and reduce:

$$R = 108.2 + 10 \log_{10} G \times 10^6/\mu f \text{ db} \qquad (2\text{-}6)$$

(5) To calculate R for magnetic fields, substitute in the previously given formulas and reduce:

$$R_{(M)} = 20 \log_{10} \left[(0.462/r_1)\sqrt{\mu/Gf} + 0.136\ r_1\sqrt{Gf/\mu} + 0.354\right] \text{ db} \qquad (2\text{-}7)$$

(6) To calculate R for electric fields, substitute in the previously given formulas and reduce:

$$R_{(E)} = 353.6 + 10 \log_{10} G/f^3\mu r_1^2 \text{ db} \qquad (2\text{-}8)$$

(7) To calculate A:

$$A = 3.338 \times 10^{-3} \times t \sqrt{Gf\mu} \text{ db} \qquad (2\text{-}9)$$

(8) To calculate B:

$$B = 20 \log_{10} \left| 1 - \left[(Z_s - Z_w)/(Z_s + Z_w)\right]^2 \exp\left[-2(\alpha + j\beta)T\right] \right|$$

$$= 20 \log_{10} \left| 1 - \left[(Z_s - Z_w)/(Z_s + Z_w)\right]^2 \times 10^{-A/10} \right. \qquad (2\text{-}10)$$

$$\left. \times (\cos 7.68 \times 10^{-4}\ t\sqrt{Gf\mu} - j \sin 7.68 \times 10^{-4} t\sqrt{Gf\mu}) \right| \text{ db}$$

If A is more than 10 db, B becomes negligible.

b. <u>Absorption Loss Calculations.</u> Absorption losses are calculated using equation 2-9:

$$A = 3.34 \times 10^{-3}\ t\ \sqrt{Gf\mu}\ \text{db}$$

As an illustration of the relative magnitude of A, assume a frequency of 1 mc for copper; then A = 3.34t. Expressed in another way, the attenuation is equal to 3.34 db per mil thickness of copper. At 10 kc, A becomes 0.334 db per mil thickness, indicating that copper becomes a poor attenuator for an rf field at low frequencies. If a permeable material is used, A becomes larger. It is best to use iron or soft steel for shielding at low frequencies.

Tables 2-3 and 2-4 present values of absorption loss for different metals. The absorption, A, for 10 mils of copper at 1 mc would be 10 times the loss for 1 mil, or 10 x 3.34 db = 33.4 db. This loss, A, is a function of the frequency (f) and increases as f increases. Satisfactory metal thickness for shielding purposes at 150 kc will also be satisfactory at higher frequencies for the same shielding purpose. The reflection loss presented to a magnetic field (equation 2-7) also increases as f increases. Comparisons of various metals and magnetic materials are given in table 2-4 as an aid in determining the type of shield required for a particular application. Note that the absorption loss of Hypernick, at 150 kc, is 88.5 db per mil. To determine the SE of Hypernick, it is necessary to multiply the absorption loss by the thickness, in mils, of the Hypernick used and then add the reflection loss. It is cautioned, however, that the high permeability is useful only if the incident field is not of sufficient intensity to saturate the metal.

c. <u>Reflection Loss Calculations.</u>

(1) The reflection loss to an electric field source that exists approximately 12 inches from the shield can be calculated using equation 2-8:

$$R_{(E)} = 353.6 + 10 \log_{10} G/f^3\mu r_1^2\ \text{db}$$

TABLE 2-3. ABSORPTION LOSS OF SOLID COPPER, ALUMINUM, AND IRON SHIELDS AT 60 CPS TO 10,000 MC

| Frequency | Copper | | Aluminum | | Iron | | Absorption Loss (db/mil) | | |
|---|---|---|---|---|---|---|---|---|---|
| | G | μ | G | μ | G | μ[a] | Copper | Aluminum | Iron |
| 60 cps | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 0.03 | 0.02 | 0.33 |
| 1000 cps | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 0.11 | 0.08 | 1.37 |
| 10 kc | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 0.33 | 0.26 | 4.35 |
| 150 kc | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 1.29 | 1.0 | 16.9 |
| 1 mc | 1 | 1 | 0.61 | 1 | 0.17 | 700 | 3.34 | 2.6 | 36.3 |
| 15 mc | 1 | 1 | 0.61 | 1 | 0.17 | 400 | 12.9 | 10 | 106.0 |
| 100 mc | 1 | 1 | 0.61 | 1 | 0.17 | 100 | 33.4 | 26 | 137.0 |
| 1500 mc | 1 | 1 | 0.61 | 1 | 0.17 | 10 | 129.0 | 100.0 | 168.0 |
| 10,000 mc | 1 | 1 | 0.61 | 1 | 0.17 | 1 | 334.0 | 260.0 | 137.0 |

[a] Other values of μ for iron are: 3 mc, 600; 10 mc, 500; and 1000 mc, 50

TABLE 2-4. ABSORPTION LOSS OF METALS AT 150 KC

| Metal | G<br>Relative<br>Conductivity | μ<br>Relative<br>Permeability<br>(at 150 kc) | Absorption Loss<br>(at 150 kc, db/mil) |
|---|---|---|---|
| Silver | 1.05 | 1 | 1.32 |
| Copper, annealed | 1.00 | 1 | 1.29 |
| Copper, hard drawn | 0.97 | 1 | 1.26 |
| Gold | 0.70 | 1 | 1.08 |
| Aluminum | 0.61 | 1 | 1.01 |
| Magnesium | 0.38 | 1 | 0.79 |
| Zinc | 0.29 | 1 | 0.70 |
| Brass | 0.26 | 1 | 0.66 |
| Cadmium | 0.23 | 1 | 0.62 |
| Nickel | 0.20 | 1 | 0.58 |
| Phosphor-bronze | 0.18 | 1 | 0.55 |
| Iron | 0.17 | 1000 | 16.9 |
| Tin | 0.15 | 1 | 0.50 |
| Steel, SAW 1045 | 0.10 | 1000 | 12.9 |
| Beryllium | 0.10 | 1 | 0.41 |
| Lead | 0.08 | 1 | 0.36 |
| Hypernick | 0.06 | 80,000 | 88.5* |
| Monel | 0.04 | 1 | 0.26 |
| Mu-Metal | 0.03 | 80,000 | 63.2* |
| Permalloy | 0.03 | 80,000 | 63.2* |
| Stainless steel | 0.02 | 1000 | 5.7 |

*Obtainable only if the incident field does not saturate the metal

For copper at 1 mc:

$$R_{(E)} = 353.6 + 10 \log_{10} \frac{1}{(10^6)^3 \times 1 \times (12)^2}$$

$$= 353.6 + 10\ (-20.16) = 152 \text{ db}$$

This value can be seen in the column for copper at 1 mc in table 2-5, where reflection loss in an electric field is listed for copper, aluminum, and iron. For a comparison of the shielding effectiveness of copper, aluminum, and steel in electric fields, compare 10 mils of each (table 2-6). Note that the absorption loss for 10 mils of copper at 1 mc is obtained from table 2-3 as 10 mils x $\frac{3.34 \text{ db}}{\text{mil}}$ = 33.4 db. For aluminum: 10 mils x $\frac{2.6 \text{ db}}{\text{mil}}$ = 26 db; and for iron: 10 mils x $\frac{36.3 \text{ db}}{\text{mil}}$ = 363 db

(2) The reflection loss to a magnetic field source that exists approximately 12 inches from the shield can be calculated using equation 2-7. For 10 mils of aluminum at 1 mc:

$$R_{(M)} = 20 \log_{10} \left( \frac{0.462}{12} \sqrt{\frac{1}{.61 \times 10^6}} + 0.136\ (12) \sqrt{\frac{.61 \times 10^6}{1}} + 0.354 \right) = 62 \text{ db} \qquad (2\text{-}11)$$

This value can be seen in the column for aluminum at 1 mc in table 2-7, where reflection losses in a magnetic field are listed for copper, aluminum, and iron. The reflection losses in a plane wave field are listed in table 2-8 for copper and iron. The comparative shielding effectiveness of copper, aluminum, and iron in magnetic fields is summarized in table 2-9. Similarly, the comparative shielding effectiveness of copper and iron for plane waves is summarized in table 2-10. The curves of absorption and reflection loss for copper and iron are shown on figures 2-24 and 2-25. The reflection loss for copper and iron in an electric, magnetic, and plane wave

TABLE 2-5. REFLECTION LOSS IN ELECTRIC FIELD (WAVE IMPEDANCE MUCH GREATER THAN 377 OHMS) OF SOLID COPPER, ALUMINUM, AND IRON SHIELDS FOR SIGNAL SOURCE 12 INCHES[a] FROM SHIELD AT 60 CPS TO 10,000 MC

| Frequency | Copper | | Aluminum | | Iron | | DB Loss[b] | |
|---|---|---|---|---|---|---|---|---|
| | G | $\mu$ | G | $\mu$ | G | $\mu$ | Copper | Aluminum |
| 60 cps | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 279 | --- |
| 1000 cps | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 242 | --- |
| 10 kc | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 212 | --- |
| 150 kc | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 177 | 175 |
| 1 mc | 1 | 1 | 0.61 | 1 | 0.17 | 700 | 152 | 150 |
| 15 mc | 1 | 1 | 0.61 | 1 | 0.17 | 400 | 117 | 115 |
| 100 mc | 1 | 1 | 0.61 | 1 | 0.17 | 100 | 92 | 90 |
| 1500 mc | 1 | 1 | 0.61 | 1 | 0.17 | 10 | c | --- |
| 10,000 mc | 1 | 1 | 0.61 | 1 | 0.17 | 1 | c | --- |

[a]For distances much greater or smaller than 12 inches, recalculate the reflection loss by using equations 2-6 or 2-8.

[b]If penetration loss is less than 10 db, total reflection loss must be corrected by use of B-factor.

[c]At these frequencies, the fields approach plane waves with an impedance of 377 ohms; see table 2-8 for plane waves.

TABLE 2-6. SHIELDING EFFECTIVENESS IN ELECTRIC FIELD (WAVE IMPEDANCE MUCH GREATER THAN 377 OHMS) OF SOLID COPPER, ALUMINUM AND IRON SHIELDS FOR SIGNAL SOURCE 12 INCHES FROM THE SHIELD AT 0.15 MC TO 100 MC

| Frequency (mc) | Copper (10 mils) A + R = SE (db) (db) (db) | Aluminum (10 mils) A + R = SE (db) (db) (db) | Iron (10 mils) A + R = SE (db) (db) (db) |
|---|---|---|---|
| 0.15 | 13 + 176 = 189 | 10 + 175 = 185 | 169 + 139 = 308 |
| 1.0 | 33 + 152 = 185 | 26 + 150 = 176 | 363 + 116 = 479 |
| 15.0 | 129 + 116 = 245 | 100 + 115 = 215 | 1060 + 83 = 1143 |
| 100 | 334 + 92 = 426 | 260 + 90 = 350 | 1370 + 64 = 1434 |

field are shown on figure 2-24. The curves are plotted for signal sources that are one foot from the shield. As frequency increases, the one-foot distance becomes a greater portion of a wave length; as the electrical distance from the source increases, the radiation approaches a plane wave. The three curves, therefore, converge as frequency increases. Even for the poorest reflection loss curve for iron at 10 kc, A is 4.35 db per mil of thickness. A shield of iron 0.03 inch thick at this frequency would therefore have a theoretical shielding effectiveness of more than 130 db.

TABLE 2-7. REFLECTION LOSS IN MAGNETIC FIELD (WAVE IMPEDANCE MUCH SMALLER THAN 377 OHMS) OF SOLID COPPER, ALUMINUM, AND IRON SHIELDS FOR SIGNAL SOURCE 12 INCHES[a] FROM THE SHIELD AT 60 CPS TO 10,000 MC

| Frequency | Copper | | Aluminum | | Iron | | DB Loss[b] | | |
|---|---|---|---|---|---|---|---|---|---|
| | G | μ | G | μ | G | μ | Copper | Aluminum | Iron |
| 60 cps | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 22 | -- | -1 |
| 1000 cps | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 34 | -- | 10 |
| 10 kc | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 44 | -- | 6 |
| 150 kc | 1 | 1 | 0.61 | 1 | 0.17 | 1000 | 56 | 54 | 19 |
| 1 mc | 1 | 1 | 0.61 | 1 | 0.17 | 700 | 64 | 62 | 28 |
| 15 mc | 1 | 1 | 0.61 | 1 | 0.17 | 400 | 76 | 74 | 42 |
| 100 mc | 1 | 1 | 0.61 | 1 | 0.17 | 100 | 84 | 82 | 56 |
| 1500 mc | 1 | 1 | 0.61 | 1 | 0.17 | 10 | c | -- | c |
| 10,000 mc | 1 | 1 | 0.61 | 1 | 0.17 | 1 | c | -- | c |

[a]For distances much greater or smaller than 12 inches, recalculate the reflection loss using the formulas given in text.

[b]If penetration loss is less than 10 db, the total reflection loss must be corrected by use of the B-factor.

[c]At these frequencies, the fields approach 377 ohms in impedance and become plane waves; see table 2-8 for plane waves.

TABLE 2-8. REFLECTION LOSS IN PLANE WAVE FIELD (WAVE IMPEDANCE EQUAL TO 377 OHMS) OF SOLID COPPER AND IRON SHIELDS FOR SIGNAL SOURCE GREATER THAN 2 $\lambda$ FROM THE SHIELD AT 60 CPS TO 10,000 MC[a]

| Frequency | Copper | | Iron | | Copper | Iron |
|---|---|---|---|---|---|---|
| | G | $\mu$ | G | $\mu$ | (loss in db)[b] | |
| 60 cps | 1 | 1 | 0.17 | 1000 | 150 | 113 |
| 1000 cps | 1 | 1 | 0.17 | 1000 | 138 | 100 |
| 10 kc | 1 | 1 | 0.17 | 1000 | 128 | 90 |
| 150 kc | 1 | 1 | 0.17 | 1000 | 117 | 79 |
| 1 mc | 1 | 1 | 0.17 | 700 | 108 | 72 |
| 15 mc | 1 | 1 | 0.17 | 400 | 96 | 63 |
| 100 mc | 1 | 1 | 0.17 | 100 | 88 | 60 |
| 1500 mc | 1 | 1 | 0.17 | 10 | 76 | 57 |
| 10,000 mc | 1 | 1 | 0.17 | 1 | 68 | 60 |

[a]Plane waves in sufficient strength below 1 mc rarely exist in the vicinity of a shielded enclosure.

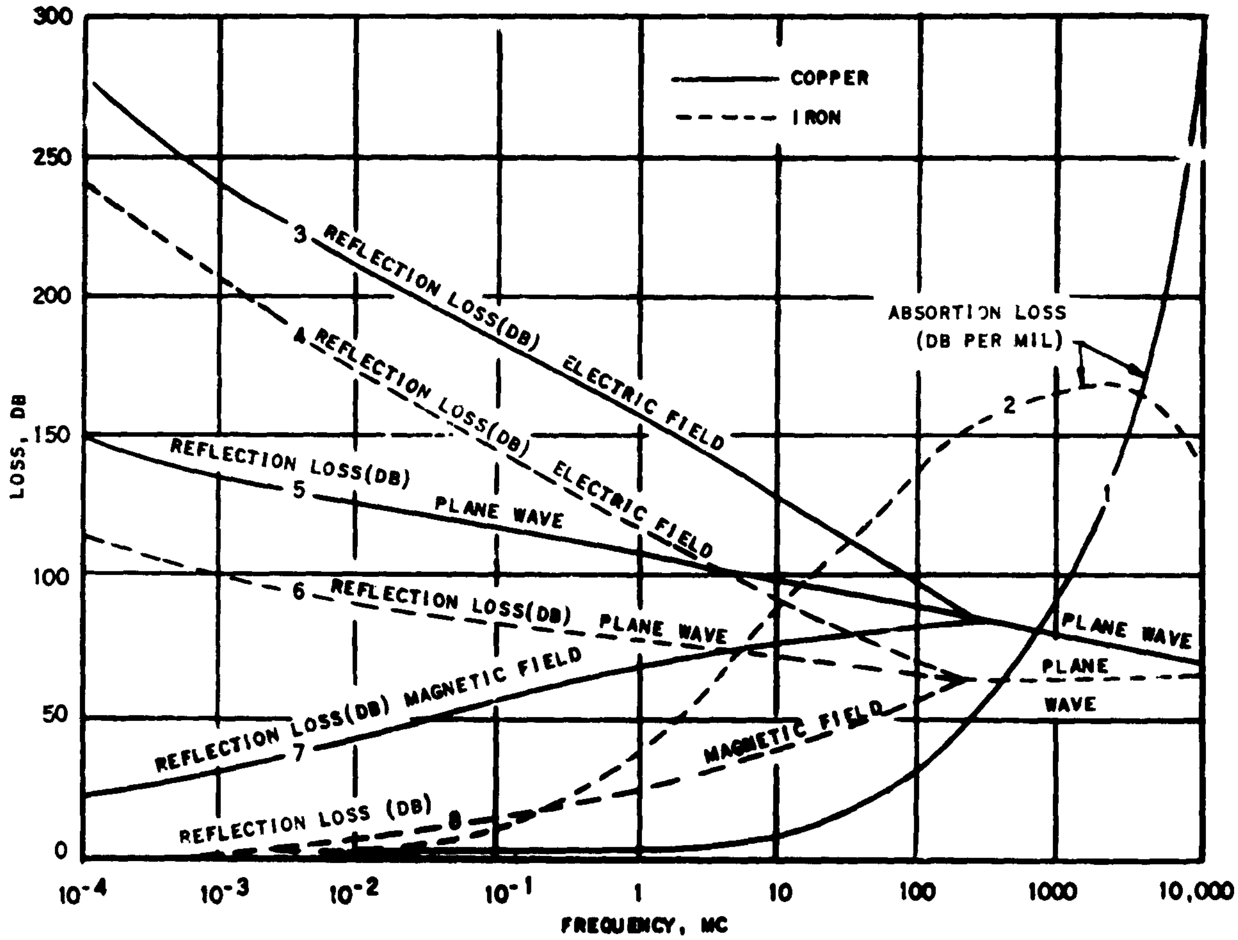
[b]If penetration loss is less than 10 db, the total reflection loss has to be corrected by use of B-factor.

TABLE 2-9. SHIELDING EFFECTIVENESS IN MAGNETIC FIELD (WAVE IMPEDANCE MUCH SMALLER THAN 377 OHMS) OF SOLID COPPER, ALUMINUM, AND IRON SHIELDS FOR SIGNAL SOURCE 12 INCHES FROM THE SHIELD AT 150 KC TO 100 MC

| Frequency (mc) | Copper (10 mils) A (db) | + | R (db) | = | SE (db) | Aluminum (10 mils) A (db) | + | R (db) | = | SE (db) | Iron (10 mils) A (db) | + | R (db) | = | SE (db) |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 0.15 | 13 | + | 56 | = | 69 | 10 | + | 54 | = | 64 | 169 | + | 19 | = | 188 |
| 1.0 | 33 | + | 64 | = | 97 | 26 | + | 62 | = | 88 | 363 | + | 28 | = | 391 |
| 15 | 129 | + | 76 | = | 205 | 100 | + | 74 | = | 174 | 1060 | + | 42 | = | 1102 |
| 100 | 334 | + | 84 | = | 418 | 260 | + | 82 | = | 342 | 1370 | + | 56 | = | 1426 |

TABLE 2-10. SHIELDING EFFECTIVENESS IN PLANE WAVE FIELD (WAVE IMPEDANCE EQUAL TO 377 OHMS) OF SOLID COPPER AND IRON SHIELDS FOR SIGNAL SOURCE GREATER THAN 2 $\lambda$ FROM THE SHIELD AT 150 KC TO 100 MC

| Frequency (mc) | Copper (10 mils) A (db) | + | R (db) | = | SE (db) | Iron (10 mils) A (db) | + | R (db) | = | SE (db) |
|---|---|---|---|---|---|---|---|---|---|---|
| 0.15 | 13 | + | 117 | = | 130 | 169 | + | 79 | = | 248 |
| 1.0 | 33 | + | 108 | = | 141 | 363 | + | 72 | = | 435 |
| 15 | 129 | + | 96 | = | 225 | 1060 | + | 63 | = | 1123 |
| 100 | 334 | + | 88 | = | 422 | 1370 | + | 60 | = | 1430 |




1 AND 2 = ABSORPTION LOSS PER MIL THICKNESS OF METAL
3 AND 4 = REFLECTION LOSS - ELECTRIC WAVES
5 AND 6 = REFLECTION LOSS - PLANE WAVES
7 AND 8 = REFLECTION LOSS - MAGNETIC FIELDS

SHIELDING EFFECTIVENESS = REFLECTION LOSS + ABSORPTION LOSS

SE = R + A

FOR COPPER: ELECTRIC FIELD; SE = CURVE 3 + CURVE 1 × "t" (THICKNESS IN MILS)*
MAGNETIC FIELD; SE = CURVE 7 + CURVE 1 × "t" (THICKNESS IN MILS)*
PLANE WAVE; SE = CURVE 5 + CURVE 1 × "t" (THICKNESS IN MILS)*
FOR IRON: ELECTRIC FIELD; SE = CURVE 4 + CURVE 2 × "t" (THICKNESS IN MILS)*
MAGNETIC FIELD; SE = CURVE 8 + CURVE 2 × "t" (THICKNESS IN MILS)*
PLANE WAVE; SE = CURVE 6 + CURVE 2 × "t" (THICKNESS IN MILS)*

* IF THE SHIELD IS ELECTRICALLY THIN (A LESS THAN 10 DB), THEN CALCULATE THE B FACTOR AND INCLUDE IN THE SHIELDING EFFECTIVENESS EQUATION (SE = R + A + B)

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Figure 2-24. Shielding Effectiveness in Electric, Magnetic and Plane Wave Fields of Solid Copper and Iron Shields (For Signal Source 12 Inches From Shield At 100 CPS To 10 KMC

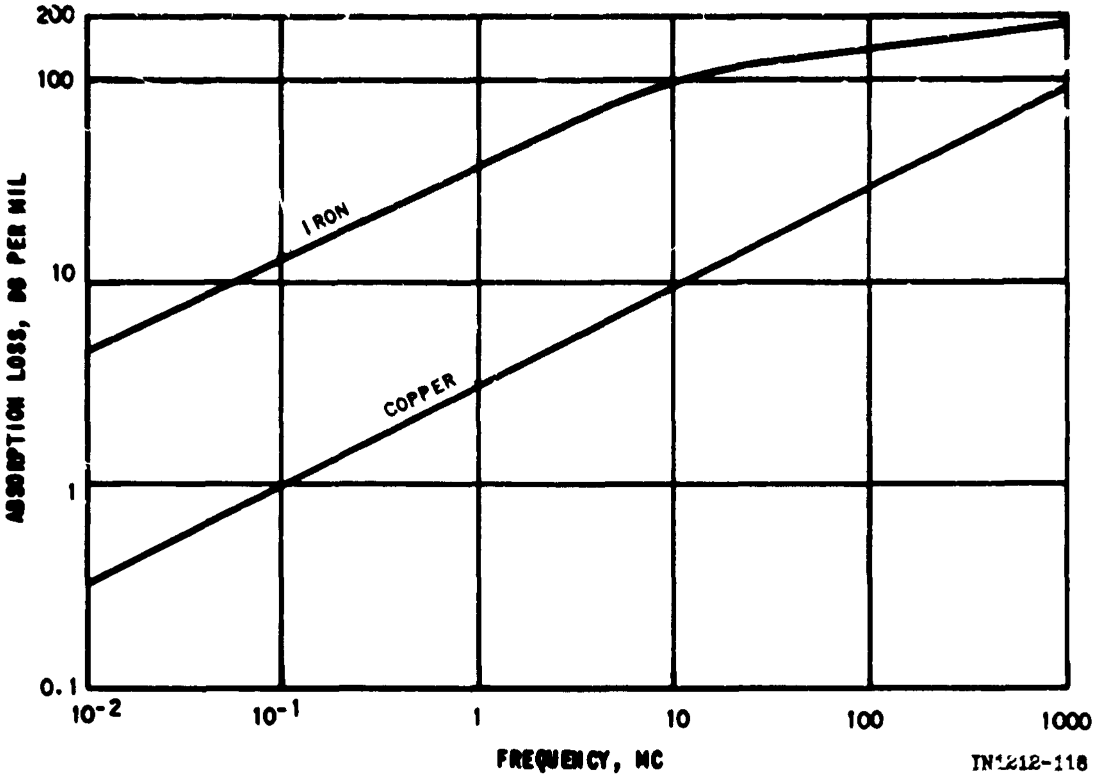



Figure 2-25. Absorption Loss for Copper and Iron, DB Per Mil

(3) To obtain shielding effectiveness values for copper and iron from figure 2-24, the following example should be used as a guide: The shielding effectiveness of copper in an electrical field is the sum of curve 3 and curve 1 multiplied by t, the thickness in mils of the copper shield. If the shield is electrically thin (A less than 10 db), then the B-factor should be calculated and included in the shielding effectiveness equation (SE = R + A + B). The B-factors for copper and iron in an electrical field, magnetic field, and for plane waves are shown in table 2-11. The ordinates for curves 1, 2, and 8, below 10 kc, on figure 2-24, are too small to be properly represented graphically; therefore, the values should be obtained from the tables. The curves were plotted for sources located a nominal distance

of 12 inches. For other distances, the losses can be recalculated. If the distance is much less than 12 inches, reflection loss to magnetic fields will be smaller, and reflection loss to electric fields greater.

(4) The results of shielding effectiveness calculations for copper, 7 mils thick, and steel, 1 mil and 50 mils thick, in electric, magnetic, and plane wave fields at a distance of 165 feet are plotted on figures 2-26, -27 and -28 and summarized in tables 2-12, -13, and -14. The shielding effectiveness is calculated by using equation 2-1: SE = R + A + B. R can be calculated for several distances for the electric, magnetic, and plane wave fields (tables 2-15, -16, and -17, respectively). The values for steel and copper are plotted on figures 2-29 and 2-30. For example, at 10 kc, for steel in an electric field at a distance of one mile, figure 2-29 shows 98 db of reflection loss. At 150 kc, for copper in a magnetic field at a distance of 24 inches, figure 2-30 shows 62 db of reflection loss. The absorption power loss, A, can be calculated for each thickness of metal. In table 2-18, the values for steel (1 mil and 50 mils thick) and copper (7 mils thick) are given. These values are plotted on figure 2-31. The B-factors are summarized in table 2-19.

(5) As a sample calculation, the shielding effectiveness of steel (1 mil thick) at 30 cps in an electric field 165 feet distant from the source is determined as follows: SE = R + A + B

where R = 203 db (from table 2-15)

A = 0.2 db (from table 2-18)

B = -27.0 db (from table 2-19)

SE = 176.2 db

TABLE 2-11. B-FACTORS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF SOLID COPPER AND IRON SHIELDS

| Shield Thickness (mils) | 60 cps | 100 cps | 1 kc | 10 kc | 100 kc | 1 mc |
|---|---|---|---|---|---|---|
| **Copper, $\mu$ = 1, G = 1, Magnetic Fields** | | | | | | |
| 1 | -22.22 | -24.31 | -28.23 | -19.61 | -10.34 | -2.61 |
| 5 | -21.30 | -22.07 | -15.83 | - 6.98 | - 0.55 | +0.14 |
| 10 | -19.23 | -18.59 | -10.37 | - 2.62 | + 0.57 | 0 |
| 20 | -15.35 | -13.77 | - 5.41 | + 0.13 | - 0.10 | |
| 30 | -12.55 | -10.76 | - 2.94 | + 0.58 | 0 | |
| 50 | - 8.88 | - 7.07 | - 0.58 | 0 | | |
| 100 | - 4.24 | - 2.74 | + 0.50 | | | |
| 200 | - 0.76 | + 0.05 | 0 | | | |
| 300 | + 0.32 | + 0.53 | | | | |
| **Copper, $\mu$ = 1, G = 1, Electric Fields and Plane Waves** | | | | | | |
| 1 | -41.52 | -39.31 | -29.38 | -19.61 | -10.33 | -2.61 |
| 5 | -27.64 | -25.46 | -15.82 | - 6.96 | - 0.55 | +0.14 |
| 10 | -21.75 | -19.61 | -10.33 | - 2.61 | + 0.57 | 0 |
| 20 | -15.99 | -13.92 | - 5.37 | + 0.14 | - 0.10 | |
| 30 | -12.73 | -10.73 | - 2.90 | + 0.58 | 0 | |
| 50 | - 8.81 | - 6.96 | - 0.55 | + 0.14 | | |
| 100 | - 4.08 | - 2.61 | + 0.51 | 0 | | |
| 200 | - 0.62 | + 0.14 | 0 | | | |
| 300 | + 0.41 | + 0.58 | | | | |
| **Iron, $\mu$ = 1000, G = 0.17, Magnetic Fields** | | | | | | |
| 1 | + 0.95 | + 1.23 | - 1.60 | - 1.83 | | |
| 5 | + 0.93 | + 0.89 | - 0.59 | 0 | | |
| 10 | + 0.78 | + 0.48 | + 0.06 | | | |
| 20 | + 0.35 | + 0.08 | 0 | | | |
| 30 | + 0.06 | - 0.06 | | | | |
| 50 | 0 | 0 | | | | |
| **Iron, $\mu$ = 1000, G = 0.17, Electric Fields and Plane Waves** | | | | | | |
| 1 | -19.53 | -17.41 | - 8.35 | - 1.31 | | |
| 5 | - 6.90 | - 5.17 | + 0.20 | 0 | | |
| 10 | - 2.56 | - 1.31 | + 0.36 | | | |
| 20 | + 0.16 | + 0.54 | 0 | | | |
| 30 | + 0.58 | + 0.42 | | | | |
| 50 | + 0.13 | 0 | | | | |

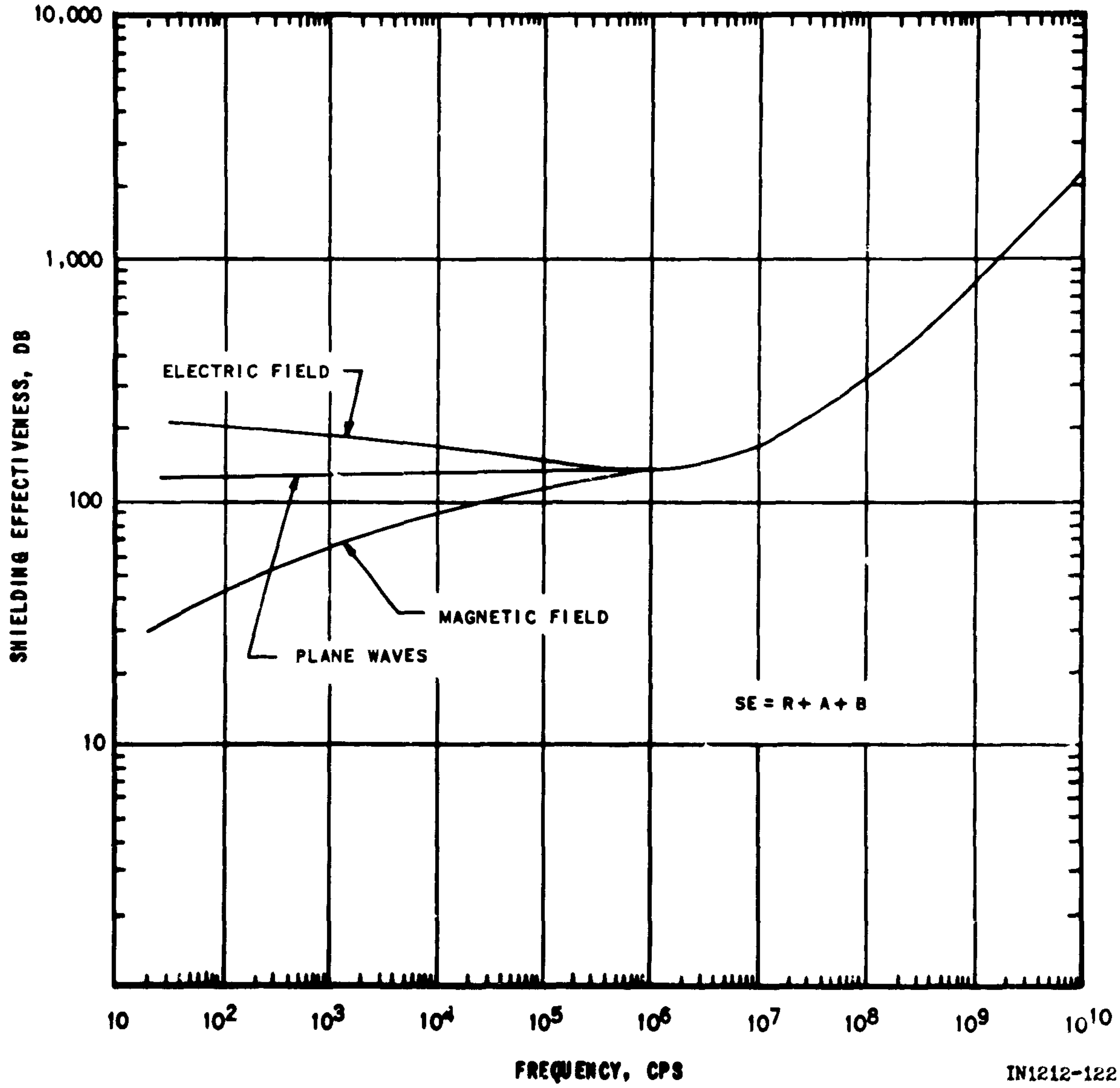



Figure 2-26. Shielding Effectiveness in Electric, Magnetic and Plane Wave Fields of Copper Shield (7 Mil Thickness) for Signal Sources 165 Feet From the Shield

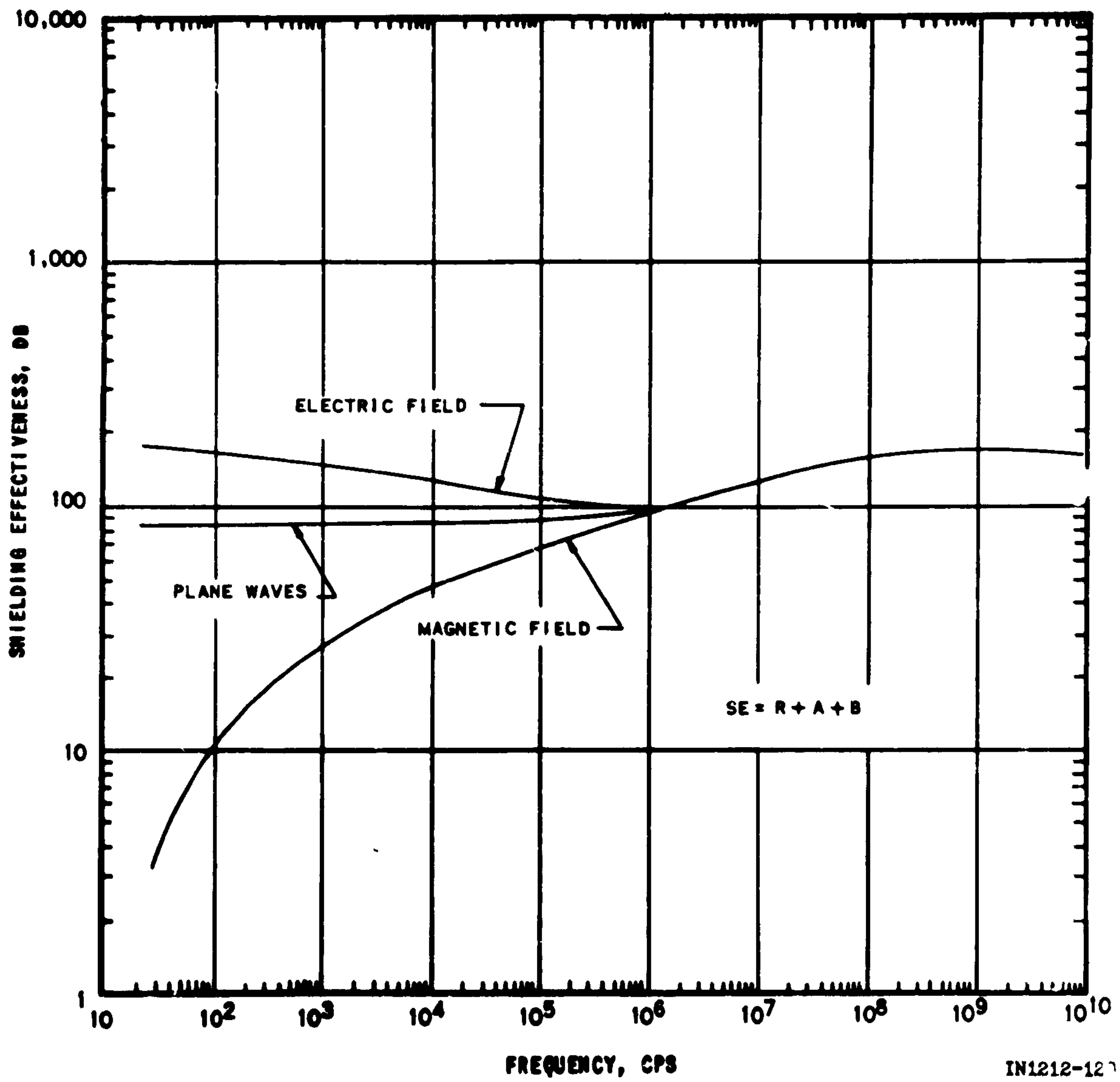



Figure 2-27. Shielding Effectiveness in Electric, Magnetic and Plane Wave Fields of Steel Shield (1 Mil Thickness) for Signal Sources 165 Feet From the Shield

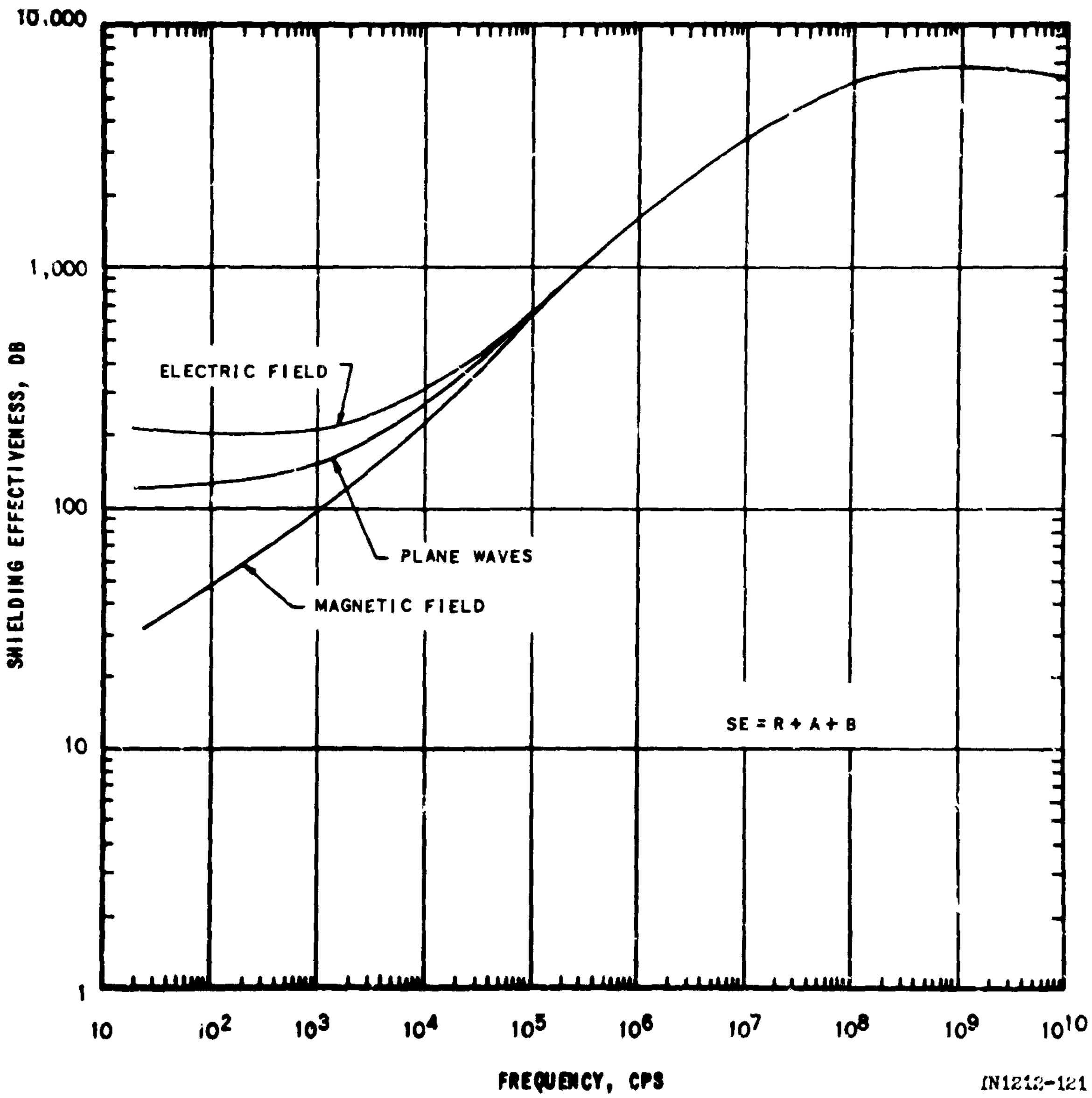



Figure 2-28. Shielding Effectiveness in Electric, Magnetic and Plane Wave Fields of Steel Shield (50 Mil Thickness) for Signal Sources 165 Feet From the Shield

TABLE 2-12. SHIELDING EFFECTIVENESS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF COPPER SHIELD (7 MIL THICKNESS) FOR SIGNAL SOURCE 165 FEET FROM THE SHIELD AT 30 CPS TO 10 KMC

| Frequency | Plane Wave (db) | Electric Field (db) | Magnetic Field (db) |
|---|---|---|---|
| 30 cps | 122 | 213 | 32 |
| 60 cps | 122 | 207 | 39 |
| 100 cps | 122 | 202 | 42 |
| 500 cps | 123 | 189 | 57 |
| 1 kc | 123 | 183 | 63 |
| 10 kc | 123 | 163 | 83 |
| 50 kc | 123 | 149 | 98 |
| 150 kc | 124 | 140 | 108 |
| 1 mc | 131 | — | — |
| 3 mc | 144 | — | — |
| 10 mc | 172 | — | — |
| 15 mc | 187 | — | — |
| 100 mc | 322 | — | — |
| 1000 mc | 818 | — | — |
| 1500 mc | 981 | — | — |
| 10 kmc | 2408 | — | — |

TABLE 2-13. SHIELDING EFFECTIVENESS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF STEEL SHIELD (1 MIL THICKNESS) FOR SIGNAL SOURCE 165 FEET FROM THE SHIELD AT 30 CPS TO 10 KMC

| Frequency | Plane Wave (db) | Electric Field (db) | Magnetic Field (db) |
|---|---|---|---|
| 30 cps | 85 | 175 | 4 |
| 60 cps | 86 | 171 | 6 |
| 100 cps | 86 | 166 | 10 |
| 500 cps | 86 | 152 | 21 |
| 1 kc | 86 | 146 | 26 |
| 10 kc | 86 | 125 | 46 |
| 50 kc | 87 | 113 | 61 |
| 150 kc | 89 | 105 | 73 |
| 1 mc | 98 | — | — |
| 3 mc | 110 | — | — |
| 10 mc | 136 | — | — |
| 15 mc | 142 | — | — |
| 100 mc | 164 | — | — |
| 1000 mc | 287 | — | — |
| 1500 mc | 186 | — | — |
| 10 kmc | 164 | — | — |

TABLE 2-14. SHIELDING EFFECTIVENESS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF STEEL SHIELD (50 MIL THICKNESS) FOR SIGNAL SOURCE 165 FEET FROM THE SHIELD AT 30 CPS TO 10 KMC

| Frequency | Plane Wave (db) | Electric Field (db) | Magnetic Field (db) |
|---|---|---|---|
| 30 cps | 121 | 211 | 31 |
| 60 cps | 123 | 208 | 39 |
| 100 cps | 125 | 205 | 46 |
| 500 cps | 138 | 204 | 73 |
| 1 kc | 151 | 211 | 91 |
| 10 kc | 249 | 289 | 210 |
| 50 kc | 455 | 481 | 430 |
| 150 kc | 725 | 741 | 709 |
| 1 mc | 1465 | — | — |
| 3 mc | 2311 | — | — |
| 10 mc | 3801 | — | — |
| 15 mc | 4140 | — | — |
| 100 mc | 5338 | — | — |
| 1000 mc | 11850 | — | — |
| 1500 mc | 6547 | — | — |
| 10 kmc | 5338 | — | — |

TABLE 2-15. TOTAL REFLECTION LOSS IN ELECTRIC FIELD (WAVE IMPEDANCE MUCH GREATER THAN 377 OHMS) AT BOTH SURFACES OF SOLID STEEL AND COPPER SHIELDS

| Frequency | DB Loss[a] | | | | | |
|---|---|---|---|---|---|---|
| | Steel | | | Copper | | |
| | Distance From Source To Shield | | | | | |
| | 24 in. | 165 ft. | 1 mile | 24 in. | 165 ft. | 1 mile |
| 30 cps | 241 | 203 | 173 | 282 | 243 | 213 |
| 60 cps | 233 | 195 | 165 | 273 | 235 | 205 |
| 100 cps | 226 | 188 | 158 | 266 | 228 | 198 |
| 500 cps | 205 | 167 | 137 | 245 | 207 | 177 |
| 1 kc | 196 | 158 | 128 | 236 | 198 | 168 |
| 10 kc | 166 | 128 | 98 | 206 | 168 | 138 |
| 50 kc | 145 | 107 | 77[b] | 185 | 147 | 117[b] |
| 150 kc | 131 | 92 | — | 171 | 132 | — |
| 1 mc | 108 | 69[b] | — | 146 | 108[b] | — |
| 3 mc | 91 | — | — | 132 | — | — |
| 10 mc | 79 | — | — | 116 | — | — |
| 15 mc | 75 | — | — | 111 | — | — |
| 100 mc | 56[b] | — | — | 86[b] | — | — |

[a]If penetration loss is less than 10 db, the total reflection loss must be corrected by use of B-factor.

[b]At these frequencies, the wave impedance approaches that of plane waves ($r > \lambda$), and the values for plane wave reflection loss should be used.

TABLE 2-16. TOTAL REFLECTION LOSS IN MAGNETIC FIELD (WAVE IMPEDANCE MUCH SMALLER THAN 377 OHMS) AT BOTH SURFACES OF SOLID STEEL AND COPPER SHIELDS

| Frequency | DB Loss[a] | | | | | |
|---|---|---|---|---|---|---|
| | Steel | | | Copper | | |
| | Distance From Source To Shield | | | | | |
| | 24 in. | 165 ft. | 1 mile | 24 in. | 165 ft. | 1 mile |
| 30 cps | -1 | 24 | 53 | 25 | 63 | 93 |
| 60 cps | -1.4 | 26 | 56 | 28 | 66 | 96 |
| 100 cps | -1.2 | 30 | 59 | 30 | 69 | 99 |
| 500 cps | 1.4 | 36 | 66 | 37 | 76 | 106 |
| 1 kc | 3.2 | 39 | 69 | 40 | 79 | 109 |
| 10 kc | 11 | 49 | 79 | 50 | 89 | 119 |
| 50 kc | 18 | 56 | 86[b] | 57 | 96 | 129[b] |
| 150 kc | 23 | 60 | — | 62 | 100 | — |
| 1 mc | 32 | 70[b] | — | 70 | 109[b] | — |
| 3 mc | 37 | — | — | 75 | — | — |
| 10 mc | 43 | — | — | 80 | — | — |
| 15 mc | 46 | — | — | 82 | — | — |
| 100 mc | 60[b] | — | — | 90[b] | — | — |

[a] If penetration loss is less than 10 db, the total reflection loss must be corrected by use of B-factor.

[b] At these frequencies, the wave impedance approaches that of plane waves ($r \gg \lambda$), and the values for plane wave reflection loss should be used.

TABLE 2-17. TOTAL REFLECTION LOSS IN PLANE WAVE FIELD (WAVE IMPEDANCE EQUALS 377 OHMS) AT BOTH SURFACES OF SOLID STEEL AND COPPER SHIELDS

| | DB Loss[a] | |
|---|---|---|
| Frequency | Steel | Copper |
| 30 cps | 113 | 153 |
| 60 cps | 110 | 150 |
| 100 cps | 108 | 148 |
| 500 cps | 101 | 141 |
| 1 kc | 98 | 138 |
| 10 kc | 88 | 128 |
| 50 kc | 81 | 121 |
| 150 kc | 76 | 116 |
| 1 mc | 70 | 108 |
| 3 mc | 66 | 103 |
| 10 mc | 61 | 98 |
| 15 mc | 60 | 96 |
| 100 mc | 58 | 88 |
| 1000 mc | 51 | 78 |
| 1500 mc | 56 | 76 |
| 10 kmc | 58 | 68 |

[a] If penetration loss is less than 10 db, the total reflection loss must be corrected by use of B-factor.

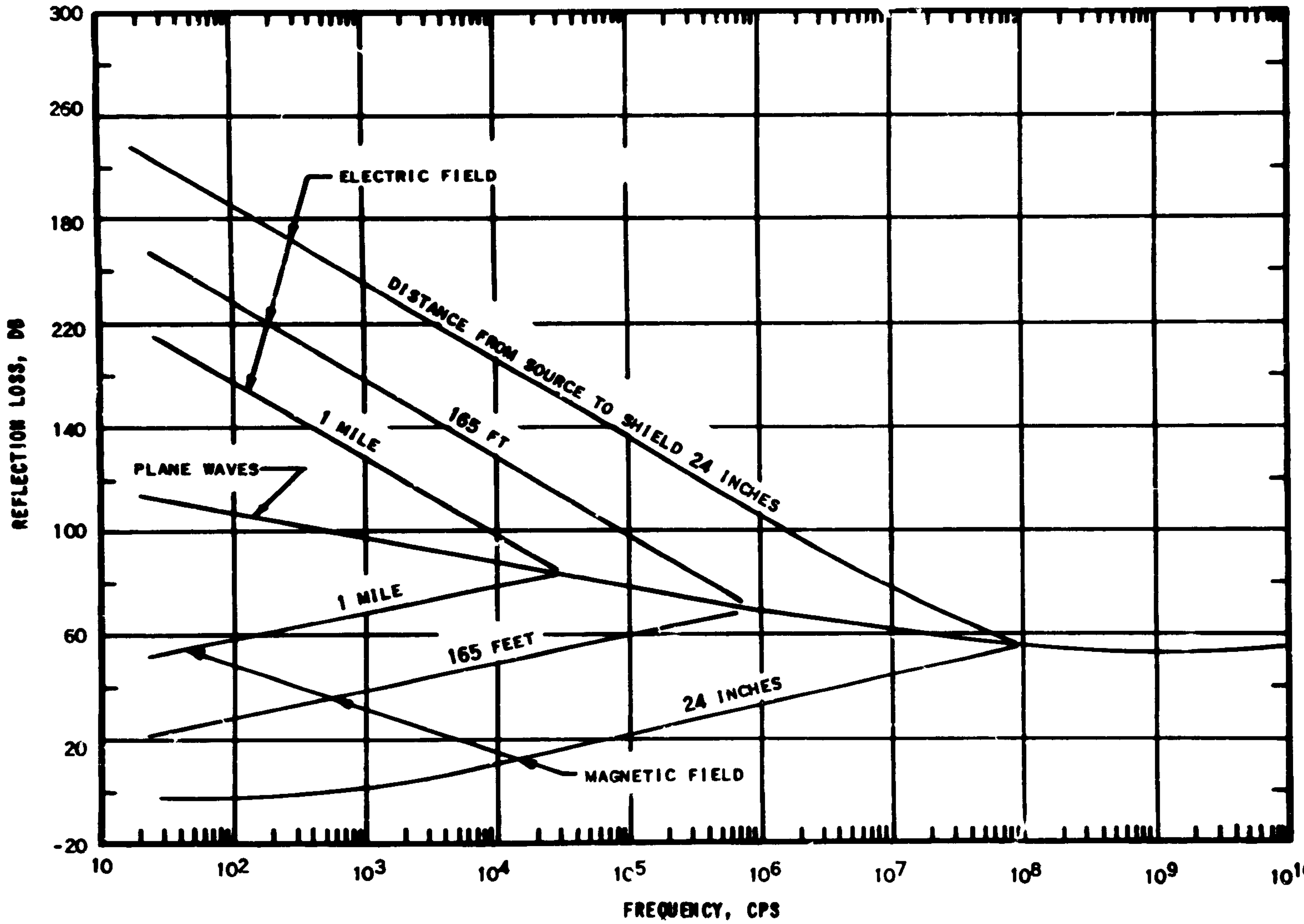



Figure 2-29. Total Reflection Loss at Both Surfaces of a Solid Steel Shield

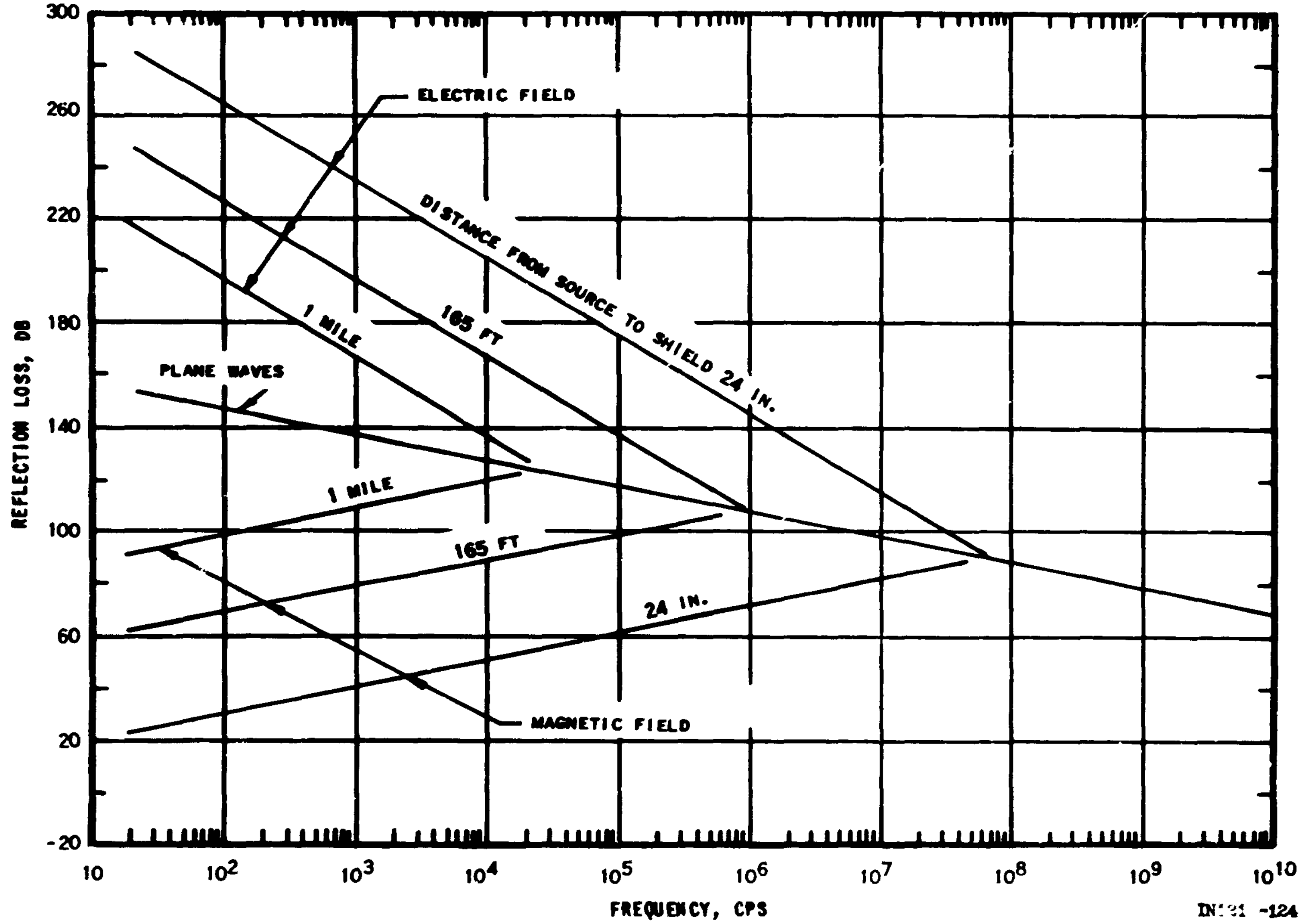



Figure 2-30. Total Reflection Loss at Both Surfaces of a Solid Copper Shield

TABLE 2-18. PENETRATION LOSS OF STEEL AND COPPER SHIELDS AT 30 CPS TO 10 KMC

| Frequency | DB Loss | | |
|---|---|---|---|
| | 1 Mil Steel | 50 Mil Steel | 7 Mil Copper |
| 30 cps | 0.2 | 9 | 0.13 |
| 60 cps | 0.3 | 13 | 0.18 |
| 100 cps | 0.3 | 17 | 0.23 |
| 500 cps | 0.7 | 37 | 0.52 |
| 1 kc | 1.0 | 53 | 0.74 |
| 10 kc | 3.2 | 161 | 2.34 |
| 50 kc | 7.5 | 374 | 5.23 |
| 150 kc | 13 | 649 | 9 |
| 1 mc | 28 | 1395 | 23 |
| 3 mc | 45 | 2245 | 40 |
| 10 mc | 75 | 3740 | 74 |
| 15 mc | 82 | 4080 | 90 |
| 100 mc | 106 | 5280 | 234 |
| 1000 mc | 236 | 11800 | 740 |
| 1500 mc | 130 | 6490 | 905 |
| 10 kmc | 106 | 5280 | 2340 |

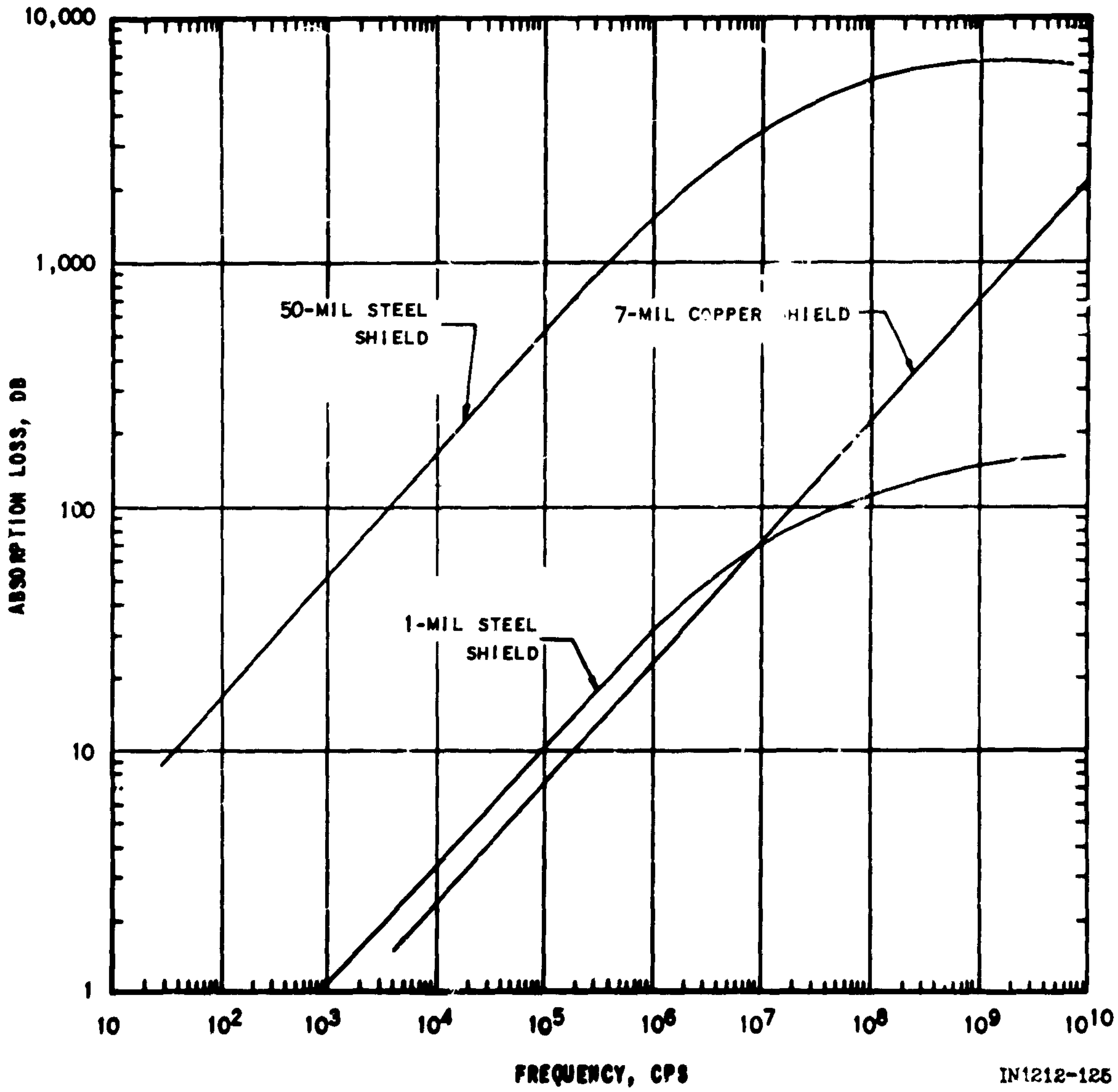



Figure 2-31. Absorption Loss for Steel and Copper Shields at 30 CPS to 10,000 MC

TABLE 2-19. CALCULATED B-FACTORS FOR STEEL AND COPPER SHIELDS FOR SIGNAL SOURCES 165 FEET FROM THE SHIELD

| | Electric Field and Plane Wave | | | | Magnetic Field | | |
|---|---|---|---|---|---|---|---|
| Frequency | 50 Mil Steel (db) | 1 Mil Steel (db) | 7 Mil Copper (db) | Frequency | 50 Mil Steel (db) | 1 Mil Steel (db) | 7 Mil Copper (db) |
| 30 cps | -1.14 | -27.0 | -30.7 | 30 cps | -1.07 | -20.0 | -30.7 |
| 60 cps | A>10 db | -24.7 | -27.9 | 60 cps | A>10 db | -20.5 | -27.9 |
| 100 cps | | -22.7 | -25.8 | 100 cps | | -19.6 | -25.8 |
| 500 cps | | -16.0 | -19.0 | 500 cps | | -15.6 | -19.0 |
| 1 kc | | -13.3 | -16.1 | 1 kc | | -13.3 | -16.1 |
| 10 kc | | - 5.6 | - 7.5 | 10 kc | | - 5.6 | - 7.5 |
| 50 kc | | - 1.7 | - 3.1 | 50 kc | | - 1.7 | - 3.1 |
| 150 kc | | A>10 db | - 1.2 | 150 kc | | A>10 db | - 1.2 |
| | | | A>10 db | | | | A>10 db |

The shielding effectiveness can be designed so that part of the burden is carried by the reflection loss and the rest by the absorption loss. The reflection loss in the electric field decreases inversely with frequency and approaches the plane wave reflection loss curve, while the reflection loss in the magnetic field increases inversely with frequency and approaches the plane wave reflection curve (fig. 2-24). The penetration loss of iron at 10 mc is about 70 db greater than the magnetic field reflection loss, but, at 100 kc, it is a few db less than the reflection loss. These factors should be fully considered when designing metal shields. For example, if shielding against electric fields at frequencies of about 10 kc is necessary, the tremendous reflection loss (over 200 db on curve 3 of figure 2-24) should be utilized rather than the absorption loss alone. To summarize, shielding effectiveness is the result of R + A + B together and not any single part. At frequencies as low as 60 cps, penetration loss and reflection loss become negligible for magnetic fields so that, for lower frequencies, very thick metallic barriers may be necessary to shield against magnetic fields.

(6) Based on the examples given in table 2-20, a metallic barrier made of iron with a thickness of 300 mils must be provided to obtain a shielding effectiveness of 100 db at 60 cps for magnetic fields. For copper and iron, reflection and penetration losses are small for magnetic fields at low frequencies. Since magnetic materials, such as Mu-metal, have high permeability at low frequencies, and therefore high absorption loss, they are more effective as shields. The resulting increase in absorption loss is obtained at the expense of a reduction in reflection loss. The following are general rules for selection of shielding materials:

TABLE 2-20. EXAMPLES FOR CALCULATING SHIELDING EFFECTIVENESS OF SOLID-METAL SHIELD

| | 10 KC —— 10 Mils | | | | | |
|---|---|---|---|---|---|---|
| | Magnetic Field | | Electric Field | | Plane Wave | |
| | Copper | Iron | Copper | Iron | Copper | Iron |
| Reflection | 44.2 | 8.0 | 212.0 | 174.0 | 128.0 | 90.5 |
| Absorption | 3.6 | 43.5 | 3.3 | 43.5 | 3.3 | 43.5 |
| B-Factor | -2.6 | 0 | -2.6 | 0 | -2.6 | 0 |
| Total Loss (db) | 45.2 | 51.5 | 212.7 | 217.5 | 128.7 | 134.0 |

| | 60 CPS —— Magnetic | | | | | |
|---|---|---|---|---|---|---|
| | 1 Mil | | 10 Mils | | 300 Mils | |
| | Copper | Iron | Copper | Iron | Copper | Iron |
| Reflection | 22.4 | -0.9 | 22.4 | -0.9 | 22.4 | -0.9 |
| Absorption | 0.03 | 0.33 | 0.26 | 3.34 | 7.80 | 100.0 |
| B-Factor | -22.2 | +0.95 | -19.2 | +0.78 | +0.32 | 0 |
| Total Loss (db) | 0.23 | 0.38 | 3.46 | 3.22 | 30.52 | 99.1 |

| | 10 KC - 30 Mils - Magnetic | | 1 KC - 10 Mils - Magnetic | |
|---|---|---|---|---|
| | Copper | Iron | Copper | Iron |
| Reflection | 44.20 | 8.0 | 34.2 | 0.9 |
| Absorption | 10.02 | 130.5 | 1.06 | 13.70 |
| B-Factor | +0.58 | 0 | -10.37 | +0.06 |
| Total Loss (db) | 54.80 | 138.5 | 24.89 | 14.66 |

| | 10 Mils —— Copper | | | | | |
|---|---|---|---|---|---|---|
| | 150 KC | | | 1 MC | | |
| | Electric | Plane Waves | Magnetic | Electric | Plane Waves | Magnetic |
| Reflection | 176.8 | 117.0 | 56.0 | 152.0 | 108.2 | 64.2 |
| Absorption | 12.9 | 12.9 | 12.9 | 33.4 | 33.4 | 33.4 |
| B-Factor | +0.5 | +0.5 | +0.5 | 0 | 0 | 0 |
| Total Loss (db) | 190.2 | 130.4 | 69.4 | 185.4 | 141.6 | 97.6 |

1) Good conductors such as copper, aluminum, and magnesium should be used for high-frequency shields to obtain the highest reflection loss

2) Magnetic materials such as iron and Mu-metal should be used for low-frequency shields to obtain the highest absorption loss

3) Any structurally sound shielding material will usually be thick enough for shielding electric fields at any frequency

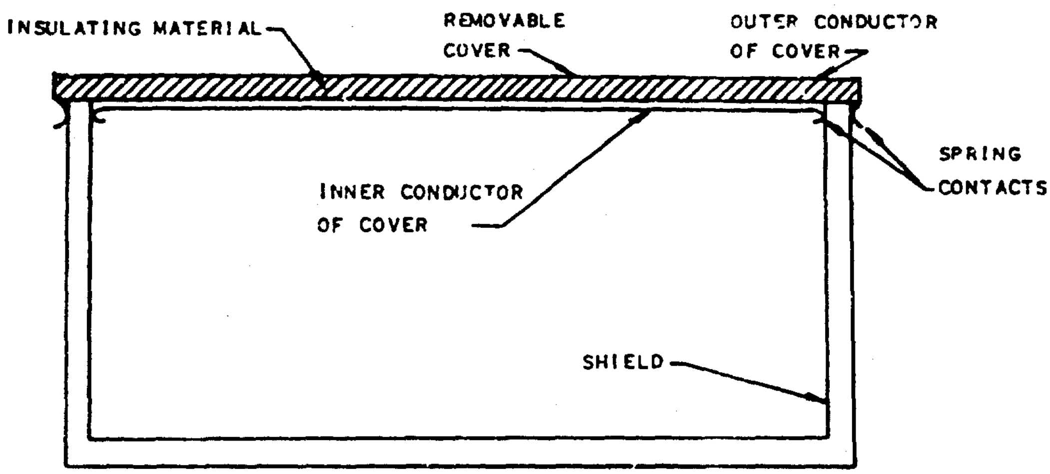
4) To provide a given degree of shielding, reference to the curves of absorption loss permit quick estimates of the required metal and thickness. In most applications, it is necessary to reduce an interference field to some specified level. This shielding requirement can be determined by measuring the interference field without the shield and comparing it with the specification limits

2-17. Multiple Shielding

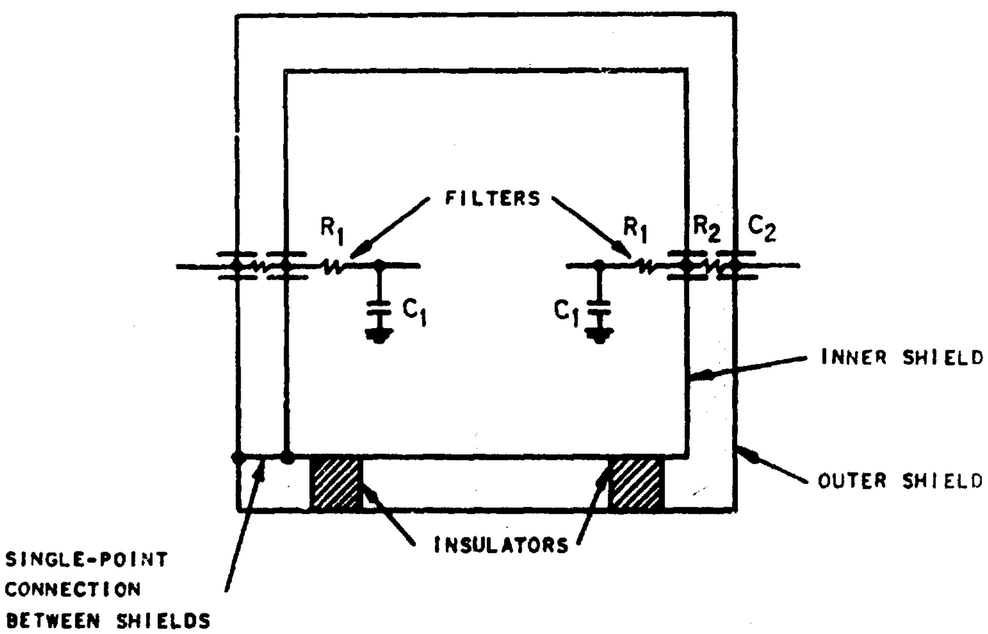
a. General. The shielding requirements necessary to protect against magnetic field transference are more difficult to achieve at low frequencies than at frequencies above several megacycles. Low-frequency shields must have either high conductivity or high permeability. In the normal power-frequency range, for example, copper must be very thick to serve as a practical magnetic shield. Mu-metal and similar type high-permeability alloys provide good shielding for low-frequency weak fields; multiple magnetic shielding is recommended for low-frequency strong fields. Power transformers and audio transformers, mounted near each other, may require multiple shielding to prevent magnetic transfer between them and to minimize interference. Good shielding effectiveness for electrical fields may be obtained with shields of high conductivity, such as copper or aluminum. Equation 2-8 shows that shielding effectiveness for electric (high-impedance) fields is infinite at zero frequency and decreases with increasing frequency. On the other hand, magnetic (low-impedance) fields are difficult to shield at low frequencies because reflection losses may approach zero for certain combinations of material and frequency. Reflection and absorption losses decrease with decreasing frequency for non-magnetic materials. At high frequencies, the shielding effectiveness is good because of reflections from the surface and rapid dissipation of the field by penetration losses. At low frequencies, it is also possible to obtain good reflection losses in magnetic materials that may be used to provide a safety factor in design. When much of the usefulness of shielding is due to reflection loss, two or more layers of metal, separated by dielectric materials and yielding multiple reflections, will provide greater shielding than the same amount of metal in a single sheet. Copper, Mu-metal, iron, conetic and netic type materials, and other metals, some with excellent electric-field reflection loss and some with excellent magnetic field absorption-loss properties, can be effectively used in combination. The results are composite antimagnetic and antielectric field shields of usable physical proportions. In many applications, it is possible to reduce shielding effectiveness requirements for the overall equipment housing by employing suppression techniques within the equipment. The recommended techniques make use of component shields, filters at the source of the undesired signal interference, partial shields, isolation of circuits by decoupling, short leads, and the ground plane as the ground return lead.

b. Multiple Shielding Applications.

(1) When shielding must be highly effective, it is customary to employ multiple shields. An effective method of handling a cabinet cover problem by multiple shielding is shown on figure 2-32A. The cover shield is in the form of a sandwich, the center of which is insulating material. The inner conducting surface of the lid makes spring contact with the inner side of the shield, while the outer conductor of the lid makes spring contact with the outer side of the shield.




**A. SANDWICH TYPE OF LID FOR A SHIELDED ENCLOSURE.**




**B. MULTIPLE SHIELDING SYSTEM PROVIDED WITH A SINGLE POINT CONNECTION BETWEEN THE SHIELDS.**

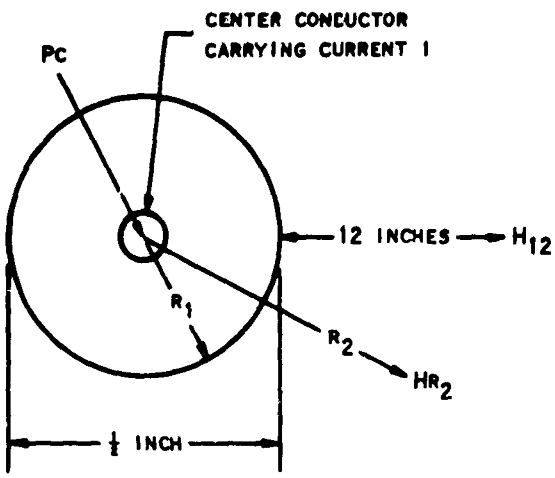
IN1212-126

Figure 2-32. Multiple Shielding Applications

Such an arrangement is several hundred times as effective as a simple, all-metal lid with spring fit, and is satisfactory under many circumstances that would otherwise require wing-nut clamping arrangements.

(2) The most important source of energy fields in a signal generator is the oscillator circuit coil; the effectiveness of the shielding system is greatly increased if the coil is enclosed in an auxiliary shield, which is placed inside the main shield. In some cases, the entire tuned circuit, or the oscillator tube and associated rf tuned circuits and chokes, are placed in a separate shield that is within the main shield. In general, refinements such as filters for leads or single-point grounding are not used in such a coil or tuned-circuit shield. If the shielding that results with such an arrangement is not adequate, the main shielding container is placed inside an outer shield. As shown on figure 2-32B, the inner shield is insulated from the outer shield except for a single connection between the two. This arrangement precludes the possibility of currents circulating around a loop completed between the shields. In such an arrangement, leads passing through both inner and outer shields are commonly provided with additional filtering located in the space between the shields. Shafts that extend from the outside to the inner compartment should be of nonconducting material to avoid introduction of additional electrical connections between the shields.

(3) To determine the thickness of the shielding required for a transmission line cable, the output of a 100-watt transmitter is fed into an antenna by a 50-ohm shielded coaxial transmission line cable. The inside diameter of the cable is 0.5 inch (fig. 2-33). Allowable magnetic field levels are: at 150 kc, up to 3000 μV/meter-kc (bandwidth); at 10 mc, up to 30 μV/




$R_1$ : RADIUS TO SHIELD OF CABLE

$R_2$ : RADIUS TO POINT IN SPACE

PC : RADIATED INTERFERENCE POWER AT CENTER OF TRANSMISSION CABLE

$H_{12}$ : MAGNETIC FIELD INTENSITY AT STODDART ANTENNA

$HR_2$ : MAGNETIC FIELD INTENSITY AT POINT IN SPACE

IN1212-127

Figure 2-33. Shielding For Transmission Cable

meter-kc (bandwidth). Both limits apply when measured with the 8-inch loop of a Stoddart NM-20 field intensity meter at a distance of 12 inches. The bandwidth of the NM-20 is 2 kc wide at 150 kc, and 6 kc wide at 10 mc. The allowable magnetic field intensities (H) at a distance of 12 inches from the shield are:

at 150 kc:

$$\frac{3000\mu V}{\text{meter-kc}} \times 2\ \text{kc} = \frac{6000\mu V}{\text{meter}}$$

$$H = \frac{6000\mu V/\text{meter}}{377\ \text{ohms}} = \frac{15.9\ \mu\text{amps}}{\text{meter}} \qquad (2\text{-}12)$$

at 10 mc:

$$\frac{30\ \mu V}{\text{meter-kc}} \times 6\ \text{kc} = \frac{180\ \mu V}{\text{meter}} \qquad (2\text{-}13)$$

$$H = \frac{180\ \mu V/\text{meter}}{377\ \text{ohms}} = \frac{477\ \mu\text{amps}}{\text{meter}}$$

$$H_{12} = \frac{1}{2\pi(r_1 + 12'')} \qquad Hr_2 = \frac{1}{2\pi r_2} \qquad (2\text{-}14)$$

$$\frac{Hr_2}{H_{12}} = \frac{1}{2\pi r_2} \qquad \frac{2\pi(r_1 + 12'')}{1} = \frac{r_1 + 12''}{r_2}$$

The allowable magnetic field intensities at a point just outside the shield ($r_2 = r_1 = 0.25$ inch) are:

at 150 kc:

$$Hr_{2(o)} = \left(\frac{0.25 + 12}{0.25}\right) \times 15.9\ \frac{\mu\text{amps}}{\text{meter}} = 780\ \frac{\mu\text{amps}}{\text{meter}} \qquad (2\text{-}15)$$

at 10 mc:

$$Hr_{2(o)} = \left(\frac{0.25 + 12}{0.25}\right) \times\ .477\ \frac{\mu\text{amps}}{\text{meter}} = 23.6\ \frac{\mu\text{amps}}{\text{meter}} \qquad (2\text{-}16)$$

The current in the transmission line conductor is:

$$100\ \text{watts} = I^2 \times 50\ \text{ohms}$$

$$I = 1.41\ \text{amps} \qquad (2\text{-}17)$$

The magnetic field level just inside the shield, 0.25 inch from the center conductor, is:

$$Hr_{2(i)} = \frac{1}{2\pi r_2} = \frac{1.41\ \text{amps} \times 10^6\ \mu\text{amps/amp}}{2\pi \times 0.25\ \text{inches}}$$

$$= .9 \times 10^6\ \frac{\mu\text{amps}}{\text{inch}} \times \frac{39.4\ \text{inches}}{\text{meter}} = 35.5 \times 10^6\ \frac{\mu\text{amps}}{\text{meter}} \qquad (2\text{-}1$$

The shielding effectiveness required at 150 kc is:

$$SE = 20 \log \frac{Hr_{2(i)}}{Hr_{2(o)}} = 20 \log \frac{35.5 \times 10^6 \ \mu\text{amps/meter}}{780 \ \mu\text{amps/meter}} \qquad (2\text{-}19)$$

$$= 93.2 \text{ db}$$

The shielding effectiveness required at 10 mc is:

$$SE = 20 \log \frac{Hr_{2(i)}}{Hr_{2(o)}} = 20 \log \frac{35.5 \times 10^6 \ \mu\text{amps/meter}}{23.6 \ \mu\text{amps/meter}} \qquad (2\text{-}20)$$

$$= 123.9 \text{ db}$$

The thickness required for a solid copper shield to reduce these interference fields to the allowable specified levels at 150 kc with 1.29 db/mil absorption loss (table 2-3) and reflection loss, 56 db, for copper (table 2-8) is as follows:

$$SE = 93.2 \text{ db} = R + A + B$$
$$= 56 \text{ db} + \frac{1.29 \text{ db}}{\text{mil}} \times t + B \qquad (2\text{-}21)$$

$$t = \frac{37.2 \text{ db}}{1.29 \text{ db/mil}} = 28.8 \text{ mils of solid copper shielding}$$

The B-factor can be neglected because the value of A is well above 10 db.

At 10 mc, the absorption loss is 9.7 db/mil (table 2-3) and the reflection loss is 72.1 db for copper (table 2-7):

$$SE = 123.9 \text{ db} = R + A + B$$
$$= 72.1 \text{ db} + 9.7 \frac{\text{db}}{\text{mil}} \times t + B \qquad (2\text{-}22)$$

$$t = \frac{51.8 \text{ db}}{9.7 \text{ db/mil}} = 5.3 \text{ mils of solid copper shielding}$$

## 2-18. Corrosion Protection

a. There are three methods for reducing corrosion and its adverse effects on shielding. The first method depends upon the use of metals which are close together in the electromotive activity series. The second method is to insulate the metals electrically, possibly with one of the organic or electrolytic finishes, and to seal the bonded joint against moisture; this method is not acceptable from the interference viewpoint. The third and preferred technique is a compromise between metal finish and interference reduction requirements. It entails selection of conductive finishes, where necessary, to eliminate the electromotive potential between dissimilar metals.

b. The most commonly used metal for military equipment is aluminum. It is chosen for its properties of low weight and high strength and is a metal with good shielding qualities. It must, however, be protected from the elements. Aluminum structural parts, which do not need to be grounded or bonded, should be anodized or given an approved chemical film finish in accordance with currently applicable specifications. A 2- to 5-minute immersion in a solution containing 5 to 10 percent chromic acid in water, and maintained at 49° to 60°C, may be used instead of anodizing; or a chemical film in accordance with currently applicable specifications may be used on parts fabricated from aluminum 1100, and aluminum alloys 3003, 5052, 6053, 6061, 6063, 7072, or equivalent. For aluminum structural parts that need to be electrically bonded or grounded, aluminum 1100, alloys 3003, 5052, 6053, 6061, 6063, 7072, or equivalent should be used, unanodized. A caustic dip, with or without water lacquer finish, is satisfactory for aluminum alloys.

c. For interference reduction purposes, mating surfaces of structures, or bonding connections between cases and mounting bases, should be free of any protective coating that would result in a higher impedance path than that provided by bare metal-to-metal contact. Those areas of a structure to be joined should be masked and kept clean before protective finishes are applied. The masked areas will then require only wire-brushing before joining. The following should be removed from the contact area:

1) Grease and oil
2) Organic finishes (zinc-chromate primers, lacquers, and enamels)
3) Inorganic finishes
4) Corrosion-preventive compounds
5) Phosphate-type finishes (Parkerizing)
6) Oxide films

The following need not be removed:

1) Cadmium plate
2) Tin plate
3) Iridite finish
4) Zinc plate without after-treatment

d. The corrosion problem is ever-present in soldering. Corroded surfaces are difficult to solder, and soldering may result in corroded surfaces unless rosin, which is noncorrosive, is used as a soldering flux. Three flux types, chloride, organic acid (waxes), and organic base, are all corrosive. They differ only in their rate of attack. Electrical joints between small, clean metal parts should be made with rosin flux, which may be conveniently applied as a core within the solder wire itself. A good soldered joint is still likely to exhibit an appreciable contact resistance.

e. Joined metals should be close together in the electromotive force series: magnesium should not be joined to brass or nickel because such a combination will cause excessive corrosion of the magnesium. If the metals are not close together in the electromotive force series, the joined metals should be of such relative sizes that the attacked metal is the more abundant. For example, iron bolts on magnesium plates are fairly satisfactory; however, aluminum rivets on brass plates are not. Another principle to observe is that joints should be kept tight and well coated to bar the entrance or exit of liquids and gases. A galvanic couple is harmless without moisture, and its corrosive power is reduced when the electrodes are coated with gas that cannot escape. Combinations such as silver and platinum, copper and monel, cadmium and steel, are known to be compatible.

f. For interference suppression purposes, Iridite No. 14 type finishes are preferable. Unlike anodizing and certain other chemical treatments, the iridite film is neither an oxide nor a phosphate; it is of a complex chromium-chromate nature and is generated by a reaction that occurs when an aluminum part is immersed in the iridite solution. In this way, the film becomes an integral part of the metal itself, rather than a superimposed film such as paint. A solution of Iridite No. 14 is applied by dip, brush, swab, or spray; it produces a clear to yellow corrosion-resistant film on aluminum and its alloys. The solution is applied at room temperature and produces a film that can be used as a clear final finish or dyed various colors. It is an ideal base for paints or lacquers and can also be used as a base for rubber bonding. The solution has a wide operating range and provides excellent corrosion resistance, easy control, and low cost. The intensity or color of the film is controlled by varying the time of treatment. A short treatment of as little as 10 seconds provides protection without changing the appearance of the aluminum; immersion for longer periods, in the order of 3 minutes, produces a yellow coating providing maximum corrosion protection. Iridite No. 14, in contrast with other protective chemical treatments for aluminum, has very little effect on electrical characteristics for either high- or low-frequency work. The coating is often used to protect abraded anodized surfaces and, at the same time, provide electrical contact to those surfaces. The iridite surface can be easily welded by the shielded arc method and can be spotwelded under limited conditions. For brush application, Iridite No. 14-2 is generally more acceptable than Iridite No. 14, since it is possible to develop the protective coating much more quickly, particularly when a high concentration is used. If the aluminum has been thoroughly deoxidized prior to treatment, Iridite No. 14 or 14-2 will be electrically conductive over the range from 60 cps to 40,000 megacycles. Compared to anodic films, iridite films exhibit extremely low electrical resistance to both ac and dc. Corrosion protection can be applied to such items as chassis shields, mounting brackets, waveguides, and connector plugs without interfering with their electrical characteristics or ground connections. Tables 2-21 and 2-22 show relative dc resistance measurements, both before and after salt-spray exposure. Figures 2-34 and 2-35 depict the equivalent rf series resistance of treated and untreated aluminum. The iridite finish satisfies Specification MIL-C-5541. A similar chromate finish, Alodyne 1200, also satisfies this specification. Alodyne 1200 is electrically compatible with iridite; both are recommended as corrosion finishes for aluminum.

2-19. Enclosure Seam Design

The final shielding effectiveness of an enclosure is mainly a function of its design rather than its construction because case openings or seams used in case construction are usually not designed to take full advantage of the capabilities of the shielding material. The design of seams requires that joints be arc welded, bolted, spot welded, or treated to produce continuous metallic contact. If a material with comparatively poor conductivity is used, the depth of material through which signals must pass should provide a shielding efficiency similar to that of the case material itself. Figure 2-36 illustrates a well-designed seam, using solder along the edges, with a conductivity 1/10 that of copper. Complex seams are openings that incorporate metallic folds or flanges as part of the design (fig. 2-37). In general, these types of seams provide better shielding efficiency than simpler types, but the efficiency obtained varies with each design.

TABLE 2-21. DC CONTACT RESISTANCE FOR ALUMINUM TEST SPECIMENS BEFORE AND AFTER SALT SPRAY EXPOSURE

| Finish | Before Salt Spray | | After Salt Spray | |
|---|---|---|---|---|
| | Number of readings | Average resistance (microhms) | Number readings | Average resistance (microhms) |
| Clean aluminum | 75 | 96 | 60 | $6085 \times 10^3$ |
| Iridite 2-minute dip | 75 | 2018 | 60 | $3366 \times 10^3$ |
| Iridite 3-minute dip | 75 | 2675 | 60 | 10,577 |
| Silver plated | 75 | All specimens less than 4 microhms | — | — |

TABLE 2-22. DC CONTACT RESISTANCE FOR ALUMINUM TEST SPECIMENS UNDER DIFFERENT PRESSURES

| Treatment | Electrical Resistance | |
|---|---|---|
| | 10-PSI Pressure | 100-PSI Pressure |
| Cleaned aluminum | 1500 microhms | 500 microhms |
| Aluminum plus Iridite No. 14 | 8000 microhms | 1900 microhms |
| Aluminum plus anodizing | Over 88 ohms | Over 88 ohms |

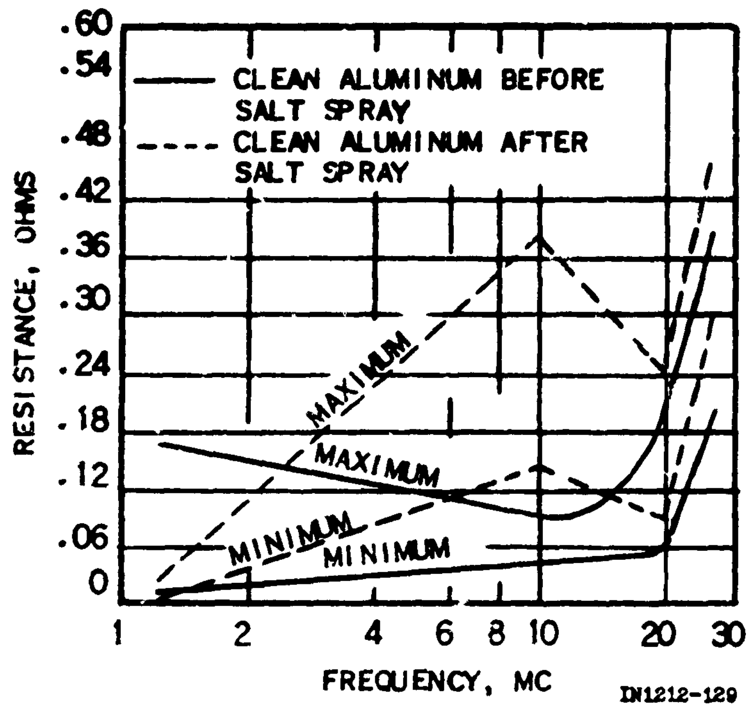



Figure 2-34. Values of Equivalent Series RF Resistance of Uncoated Aluminum Sections Before and After 64-Hour Salt Spray Test

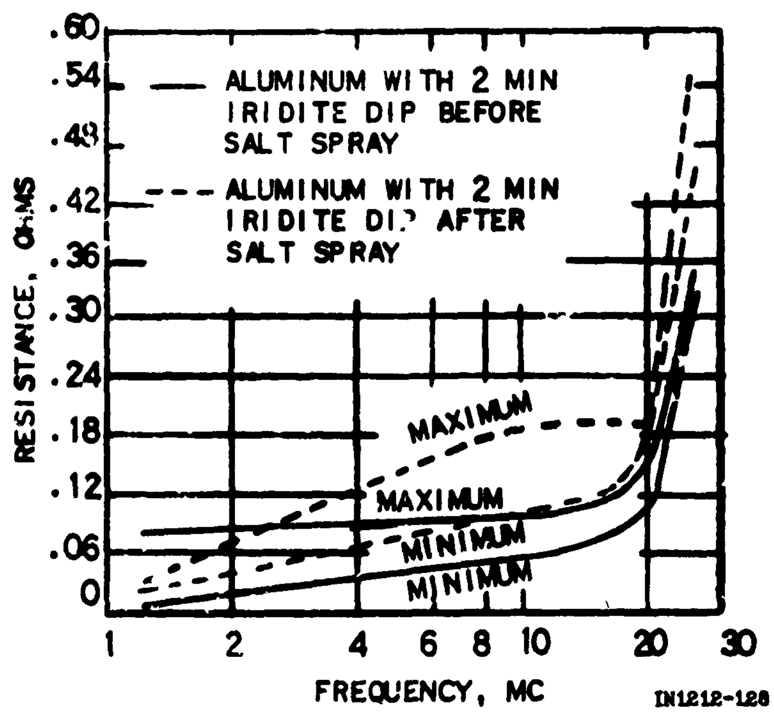



Figure 2-35. Values of Equivalent Series RF Resistance of Chromate-Treated Aluminum Sections Before and After 64-Hour Salt Spray Test

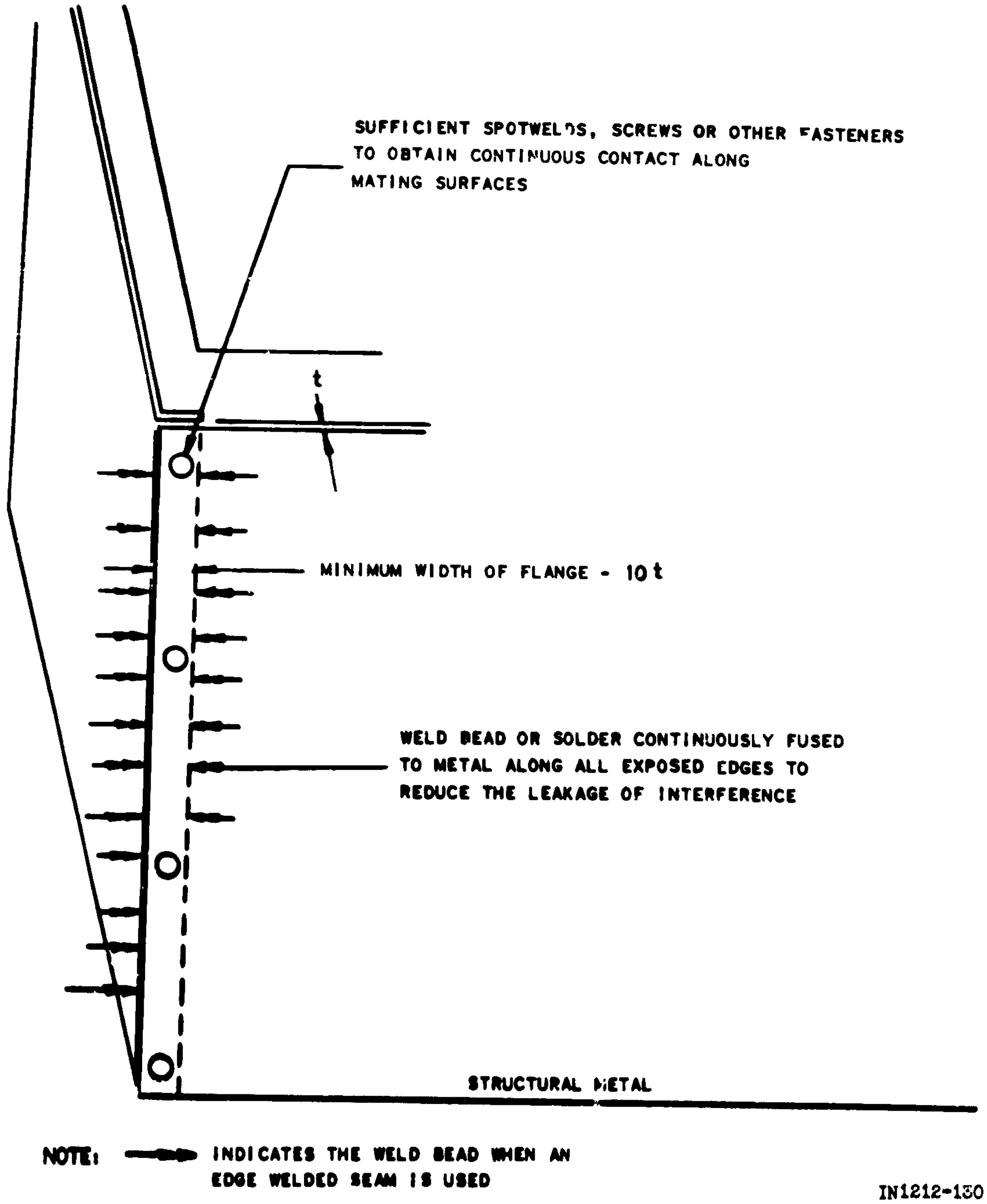



Figure 2-36. Seam Design for Minimum Interference

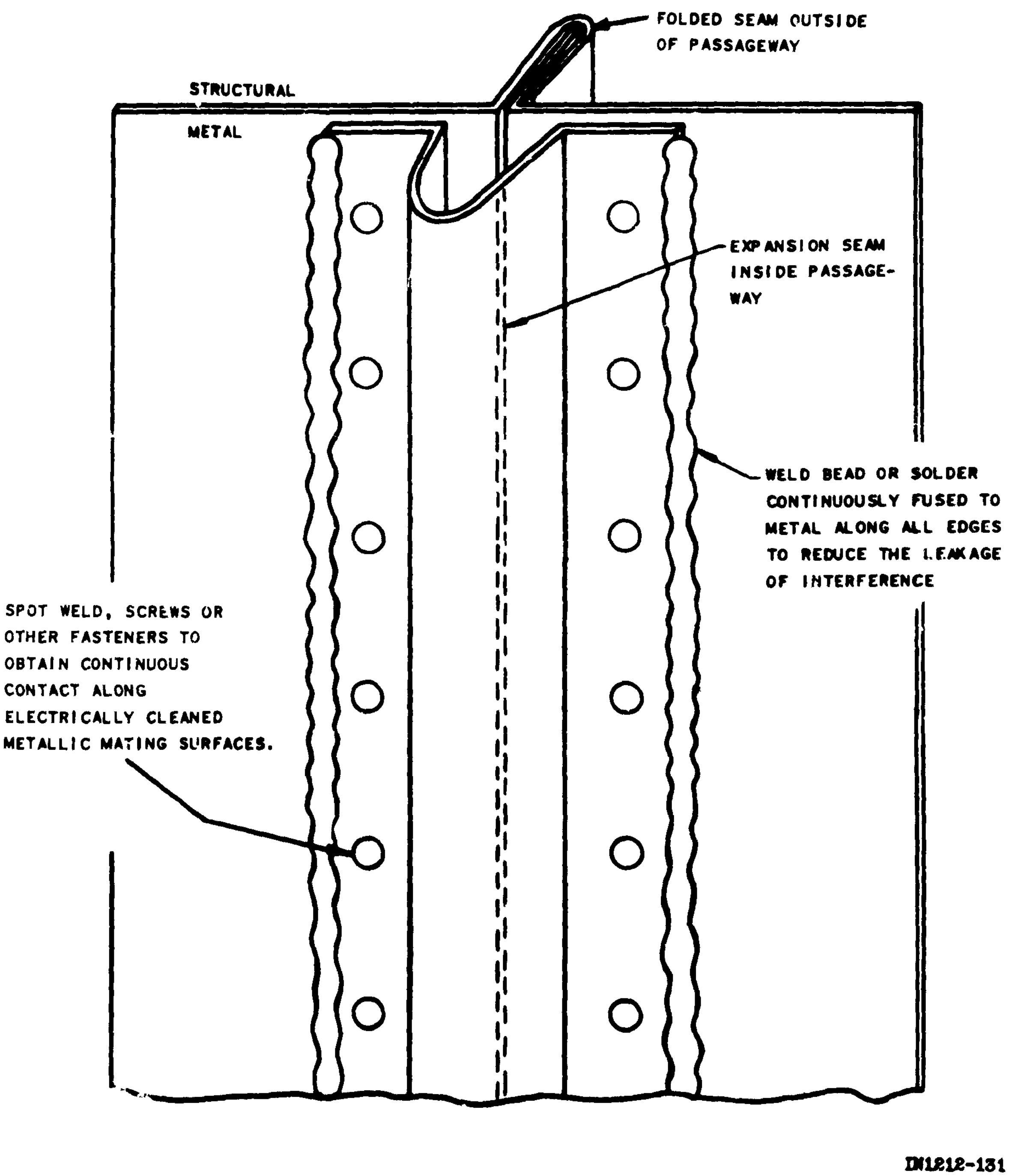



Figure 2-37. Vertical Expansion Joint, an Example of a Complex Seam

2-20. Apertures

a. General. The final determination of whether an equipment enclosure is satisfactory can be established by a radiated interference test. Such a test, when performed in accordance with military interference test specifications, requires specified antennas to receive radiated interference. The level of received radiation is a function of the radiation effectiveness of the source acting through the equipment enclosure. A single opening in the equipment enclosure will enable it to act as a relatively poor radiator; more than one opening will enable it to radiate more efficiently. A special case may exist where the energy passing through a single opening irradiates the surface of an adjacent cable shield and radiates outside of the enclosure at high efficiency. When holes in the bottom of an enclosure are for drainage of condensed moisture only, a few holes, no more than 0.25-inch in diameter, are usually sufficient. Leakage of electromagnetic energy through these holes is a small percentage of the total generated power and is negligible except in the case of high-power generators, such as radar modulators. Discretion should be exercised in determining the maximum dimensions of an aperture for a specified frequency range because interference leakage increases with frequency.

b. Shielding Effectiveness For Perforated Metal Shields.

(1) Shielding Effectiveness Formulas

The shielding effectiveness of metal shields has been treated from the viewpoint of classical transmission line theory in equation 2-1. To obtain a more rigorous expression for shielding effectiveness it is necessary to account for the following:

1) The attenuation effects of the individual shield apertures acting as many stacked waveguides beyond cutoff. For this reason, equations 2-24 and 2-25 below are used in place of equation 2-9

2) Reflection losses considering the geometry of the openings

3) Area of the openings when the test antenna is far from the shield in comparison to the distance between holes in the shield

4) Skin depth effects

5) Coupling between closely spaced openings

The shielding effectiveness, in decibels, is expressed as follows:

$$SE = A_a + R_a + B_a + K_1 + K_2 + K_3 \qquad (2\text{-}23)$$

where:

$A_a$ = aperture attenuation

$$= 27.3 \frac{D}{W} \text{ for rectangular apertures} \qquad (2\text{-}24)$$

$$= 32 \frac{D}{d} \text{ for circular apertures} \qquad (2\text{-}25)$$

D = depth of apertures in inches

W = width of rectangular apertures in inches

d = diameter of circular apertures in inches

$R_a$ = aperture reflection losses

$$= 20 \log_{10} \frac{\left|1 + k\right|^2}{4 \left|k\right|} \qquad (2\text{-}26)$$

$k = \frac{Z_s}{Z_w}$ = ratio of aperture characteristic impedance to impedance of the incident wave

= W/3.142r for rectangular apertures and magnetic fields (2-27)

= d/3.682r for c culer apertures and magnetic fields (2-28)

$= jfW\ 1.7 \times 10^{-4}$ for rectangular apertures and radiated fields (2-29)

$= jfd\ 1.47 \times 10^{-4}$ for circular apertures and radiated fields (2-30)

f = frequency in megacycles per second

r = distance from signal source to shield in inches

$B_a$ = correction factor for aperture reflections. $B_a$ becomes insignificant when $A_a$ is more than 10 db.

$$= 20 \log_{10} \left| 1 - \frac{(k-1)^2}{(k+1)^2} 10^{-\frac{A_a}{10}} \right| \tag{2-31}$$

$K_1$ = correction factor for the number of openings per unit square when the test antennas are far from the shield in comparison to the distance between holes in the shield

$$= 10 \log_{10} \frac{1}{an} \tag{2-32}$$

a = area of each hole in square inches

n = number of holes per square inch

$K_2$ = correction factor for penetration of the conductor at low frequencies

$$= -20 \log_{10} \left[ 1 + \frac{35}{p^{2.3}} \right] \tag{2-33}$$

p = ratio of the wire diameter to skin depth for screening

= ratio of the conductor width to skin depth between holes for perforated sheets. (Copper skin depth = $\frac{2.6 \times 10^{-3}}{\sqrt{f}}$ inches)

$K_3$ = correction factor for coupling between closely spaced shallow holes

$$= 20 \log_{10} \frac{1}{\tanh \left| \frac{A_a}{8.686} \right|} \qquad (2\text{-}34)$$

(2) Shielding Effectiveness Formula Derivations

(a) Attenuation calculations ($A_a$). In equation 2-23, $A_a$ represents the attenuation as the wave passes through the aperture which, for frequencies below cutoff in rectangular guides, is given as:

$$A_a = 8.68\pi \frac{D}{W} \sqrt{1 - \left(\frac{f}{f_c}\right)^2} \text{ db} \qquad (2\text{-}35)$$

where f and $f_c$ are the frequency under consideration and the cutoff frequency, respectively. It will be observed that W is always that hole dimension perpendicular to the E field. Equation 2-35, evaluated for frequencies well below cutoff, provides equation 2-24; and a similar procedure for circular guides provides equation 2-25.

(b) Reflection calculations ($R_a$). In equation 2-26, the reflection losses are calculated as a function of the ratio of the waveguide characteristic impedance below cutoff to the incident wave impedance. The characteristic impedance of a rectangular waveguide well below cutoff is:

$$Z_a = \frac{j\omega\mu W}{\pi} \qquad (2\text{-}36)$$

The impedance of the wave, emitted by a small loop source at points close to the source, compared to a wavelength, is:

$$Z_w = j\omega\mu r \qquad (2\text{-}37)$$

Taking the ratio of equation 2-36 to equation 2-37, we have:

$$k = \frac{W}{\pi r}$$

which is the equation for k given by equation 2-27 for magnetic fields. Both equations 2-36 and 2-37 are in MKS units, but, in taking the ratio, W and r may be expressed in either inches or meters. Similar procedures provide equations 2-28, -29 and -30.

(c) Corrections to reflection calculations ($B_a$). This factor is given for metal shields as:

$$B = 20 \log_{10} \left| 1 - \frac{(k - 1)^2}{(k + 1)^2} e^{-2\sigma t} \right| \qquad (2\text{-}38)$$

where $\sigma$ and t are, respectively, the complex propagation constant and the thickness of the shield in MKS units. In a waveguide below cutoff, the phase constant approaches zero and the propagation constant becomes equal to the attenuation constant, so that $e^{-2\sigma t}$ becomes equal to the reduction in signal intensity in nepers for twice the depth of the waveguide. The expression $e^{-2\sigma t}$, therefore, may be expressed in decibels and is equal to $10^{-\frac{A_a}{10}}$ which converts equation 2-38 to equation 2-31.

(d) Corrections for the number of openings that must be considered ($K_1$). It is obvious that when electromagnetic signals pass through a metal shield by penetration of openings, the amount of power transferred from one side of the shield to the other is a function of the number of openings. Not so obvious is the fact that if insertion loss tests are performed on the shield, the results will be a function of the distance between the antenna and the surface of the shield, assuming the shield to be centrally located between the antennas. If small antennas of approximately the same size as the openings, or smaller, are used for the test and are located on each side and adjacent to one of the openings, the measured shielding efficiency will be that of the opening itself. On the other hand, if the antennas are located at a considerable distance from the shield in comparison to the distance between holes in the shield, the measured shielding effectiveness will be equal to that of a single opening plus the ratio (in decibels) of the total wall area illuminated by the radiator to the total opening area located in the illuminated region. If the openings are evenly distributed, this ratio is a constant, since any change in the wall area illuminated will cause a similar change in hole area. Therefore, the minimum shielding effectiveness becomes that of the single opening plus the correction factor of equation 2-32. At intermediate points, the shielding effectiveness lies between the two values. Both values are of practical importance. The shielding effectiveness near the shield is important for protection of sensitive equipments which may be placed close to the walls (data cables, etc.). It is especially important in cases where radiation hazards may exist, so that personnel are protected who otherwise may unknowingly enter a strong field existing in the vicinity of a poorly designed seam. Calculations of shielding effectiveness in interior parts of a structure are important since, if advantage is taken of added shielding effectiveness, a considerable reduction in cost of shield construction can be achieved.

(e) Low frequency corrections for screen-type shields ($K_2$).

Numerous tests have shown that the high-frequency shielding effectiveness of screening materials can be satisfactorily approximated by assuming that the openings in the screen are equivalent in size to the openings in a flat perforated metal sheet, and that the depth of the openings is equal to the wire diameter. At low frequencies, when the skin depth becomes comparable to the radius of the wire, a considerable loss in shielding effectiveness occurs. This may be considered from the viewpoint that the apertures, as waveguides, are made wider and shorter by a skin depth. However, this approach runs into difficulty when the skin depth becomes equal to or greater than the wire radius since this is the borderline region where leakage through the metal itself must also be considered. To maintain reasonable simplicity for calculation purposes, test results for a variety of copper screen shields were plotted as a function of skin depth, and an empirical equation (2-33) was derived for the correction factor. This correction factor may also be satisfactory for perforated sheet metal, but no corroborative tests have been made.

(f) Corrections for closely spaced shallow openings ($K_3$).

When apertures in a shield are closely spaced, and the depth of the openings is small compared to the width, the shielding effectiveness has been found to be greater than otherwise would be expected. This is interpreted as being a result of coupling between adjacent holes, which becomes important when the attenuation through the openings is small. Considering two such adjacent holes subjected to an electromagnetic field, aligned as shown in figure 2-38, it appears that current induced on the conductor between the holes can flow into one side of a hole and return immediately via the adjacent hole -- in effect, merely encircling the conductor. Since the current is the same in closely spaced holes, this is equivalent to placing practically a dead short circuit at the end of each hole considered as a waveguide. The impedance of the short may be approximated for rectangular holes as the surface impedance presented by the surface of the conductor between the holes:

$$Z_L = \sqrt{\frac{j\omega\mu}{\sigma}} \cdot \frac{a}{b} \qquad (2\text{-}39)$$

where: $\sqrt{\frac{j\omega\mu}{\sigma}}$ is the intrinsic impedance of the metal (MKS units)

$\omega$ = $2\pi f$

$\mu$ = permeability of the metal = [illegible] $\times 10^{-6}$ henries/meter for copper

$\sigma$ = conductivity of the metal = $5.8 \times 10^7$ mhos/meter for copper

a = width of conductor between holes (fig. 2-38)

b = length of rectangular hole (fig. 2-38)

The characteristic aperture impedance given by equation 2-36 is $\frac{j\omega\mu W}{\pi}$ at all frequencies being considered. The ratio of $Z_L$ to $Z_a$ is therefore:

$$\frac{Z_L}{Z_a} = \sqrt{\frac{j\omega\mu}{\sigma}} \cdot \frac{a}{b} \cdot \frac{\pi}{j\omega\mu W} = \frac{a\pi}{bW} \sqrt{\frac{1}{j\omega\mu\sigma}} \qquad (2\text{-}40)$$

At frequencies as low as 10 kc, the expression under the

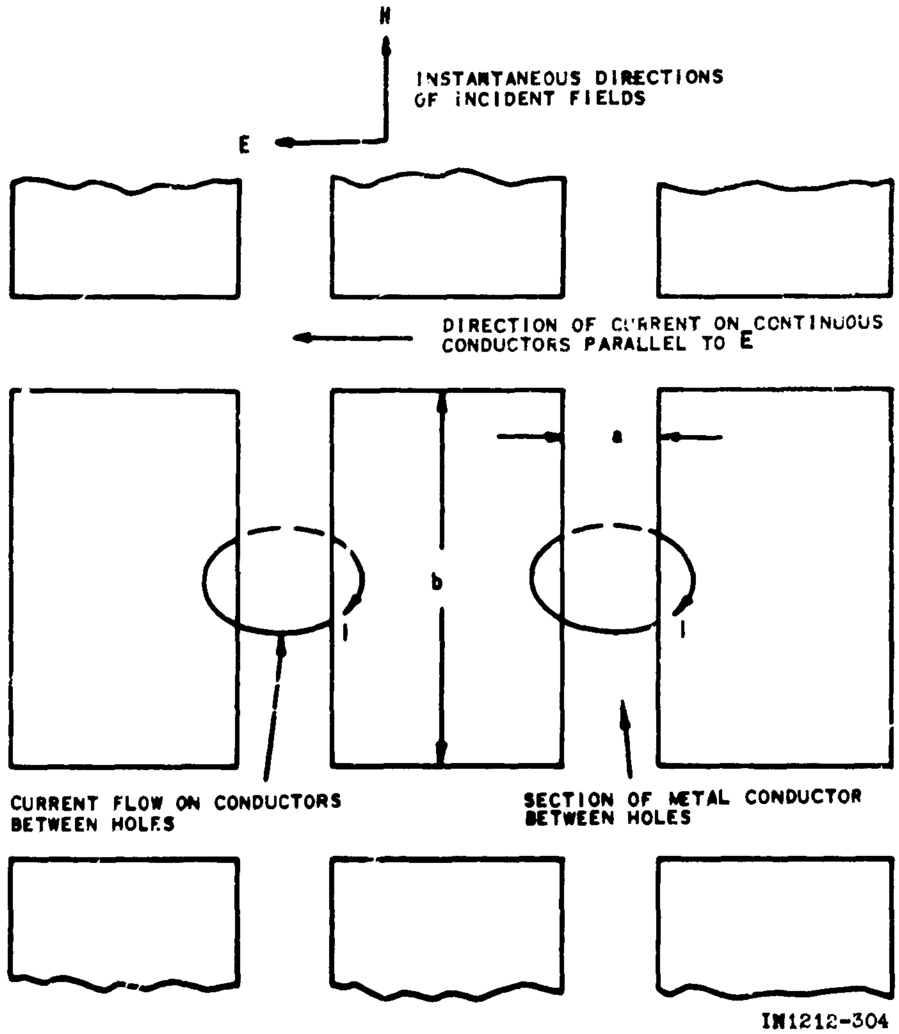



Figure 2-38. Currents Induced On Perforated Metal Sheet

radical is smaller than $10^{-6}$, showing that for all reasonable opening dimensions and any frequency of interest, $Z_L$ is much smaller than $Z_a$.

In accordance with transmission line theory, the input impedance of the guide, $Z_i$, may be calculated by:

$$Z_i = Z_a \frac{Z_L \cosh\alpha + Z_a \sinh\alpha}{Z_a \cosh\alpha + Z_L \sinh\alpha} \qquad (2\text{-}41)$$

where $\alpha$ is the waveguide attenuation in nepers

Since, for the express condition being investigated, $\alpha$ is small, and since $Z_L \ll Z_a$, eq ation 2-41 simplifies to:

$$Z_i = Z_a \frac{\sinh\alpha}{\cosh\alpha} = Z_a \tanh\alpha \qquad (2\text{-}42)$$

The ratio of the intensities of the reflected wave to the transmitted wave, when the attenuation is large, is equal to:

$$\frac{\left(Z_w + Z_a\right)^2}{4 Z_w Z_a} \qquad (2\text{-}43)$$

When the atten ation is small, the ratio becomes:

$$\frac{\left(Z_w + Z_i\right)^2}{4 Z_w Z_i} \qquad (2\text{-}44)$$

For all practical purposes, the wave impedance $Z_w$ is always much larger than either $Z_a$ or $Z_i$, and the ratio of equations 2-44 to 2-43 reduces to:

$$\frac{s'}{Z_i} = \frac{1}{\tanh\alpha} \qquad (2\text{-}45)$$

Equation 2-45, in de ibels, provides the correction factor for closely spaced sha w holes as given by equation 2-34. It will be observed that, s ce $Z_a$ is always larger than $Z_i$ for $A_a$ equal to 10 db or less, the correction factor is always positive and increases the shielding effectiveness.

c. Screening.

(1) An equipment enclosure that requires inlet and/or outlet apertures for ventilation or pressurization should be designed with a screen or a series of honeycomb tube ducts (designed to act as waveguide below cutoff devices) placed over the ventilation apertures. Although louvered openings are generally used for

cooling air circulation, they are extremely poor for rf integrity because of their long, narrow gaps. In descending order of attenuation properties, the following materials should be used: honeycomb-type ventilation panels, perforated metal sheet, woven metal mesh, and knitted metal mesh. The honeycomb material also has the advantage of low air resistance (figs. 2-39 and 2-40). The honeycomb ventilation material is shown on figure 2-41.

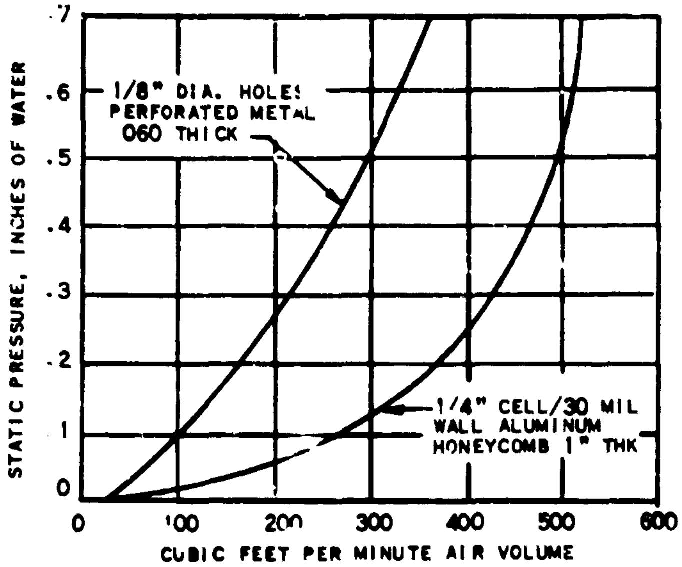



Figure 2-39. Air Impedance of Perforated Metal and Honeycomb

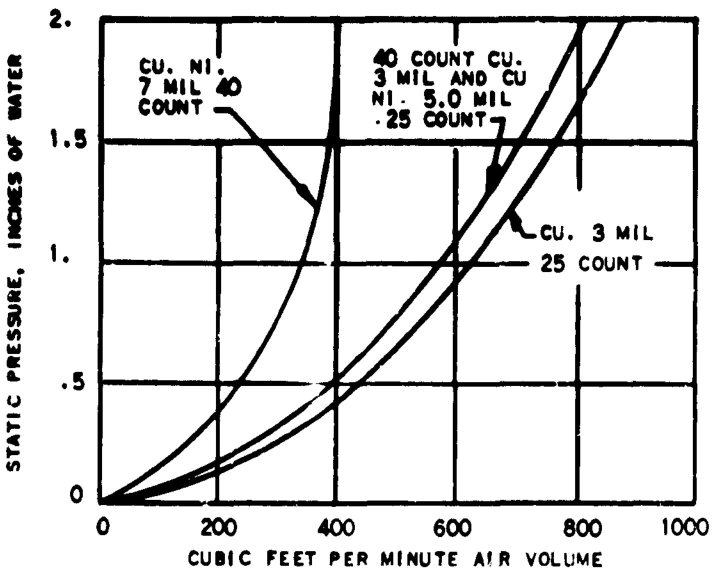



Figure 2-40. Air Impedance of Copper and Nickel Mesh

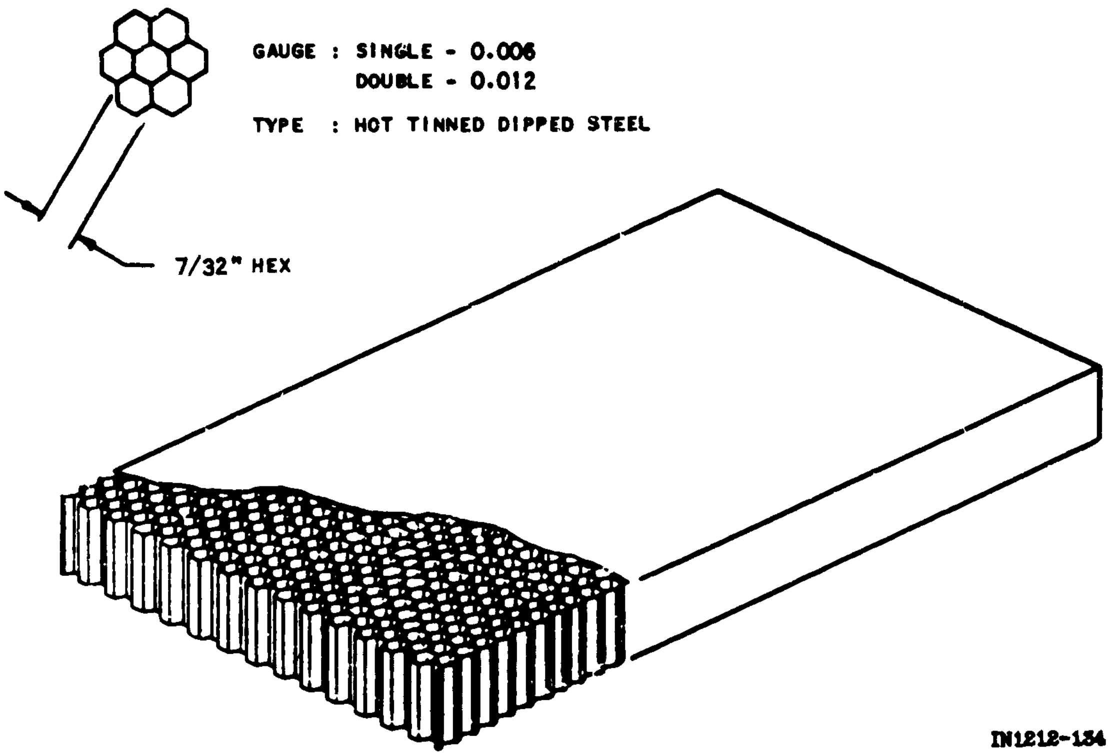



Figure 2-41. Honeycomb Type Ventilation Panel

(2) The use of a single or double layer of copper or brass screen, of No. 16 or 22 gauge wire, having openings no greater than 1/16 by 1/16 inch is recommended. A mesh less than 18 by 18 (wires to the inch) should not be used. The mesh wire diameter should be a minimum of 0.025 inch (No. 22 AWG). If more than a nominal 50 db of attenuation is required, the screening should have holes no larger than those in a 22 by 22 mesh made of 15 mil copper wires.

(3) The attenuation of an electromagnetic wave by a mesh is considerably less than that afforded by a solid metal screen. The principal shielding action of a mesh is due to reflection. Tests have shown that mesn with 50 per cent open area and 60 or more strands per wave length introduces a reflection loss very nearly equal to that of a solid sheet of the same material.

The mesh construction should have individual strands permanently joined at points of intersection by a fusing process so that permanent electrical contact is made and oxidation does not reduce shielding effectiveness. A screen of this construction will be very effective for shielding against electric (high-impedance) fields at low frequencies because the losses will be primarily caused by reflection. Installation can be made by connecting a screen around the periphery of an opening (figs. 2-42 and 2-43). The results of shielding effectiveness measurements for screening are shown on figures 2-44 through 2-49. Table 2-23 lists mesh, wire, and aperture sizes for various screening materials.

(4) The screening materials given on table 2-24 and on figures 2-50 and 2-51 exhibit low-shielding effectiveness -- probably because of crossover discontinuities in mesh. A comparison between No. 16 aluminum mesh and No. 12 copper mesh (fig. 2-50) reveals a 5 db differential, which is comparable to the 60 mils perforated steel shown on figure 2-51. All are in the 50-db range, with the perforated material having a greater percentage of open area and, therefore, less resistance to air flow. Because mesh, for effective shielding, rarely has more than an open area of 50 percent, the size of apertures must be correspondingly increased for effective ventilation. Mesh should be easily removable; it should be attached with screws or bolts. These should be in sufficient number to ensure high-pressure contact along a continuous line completely around the edge. Contact surfaces should be thoroughly cleaned each time the mesh is removed.

(5) In the 40-db range, the 30-mil, 1/4-inch spacing galvanized steel mesh and the 0.037-inch aluminum perforated sheet (fig. 2-51) are comparable to No. 10, 18 mils monel mesh (fig. 2-50). The galvanized mesh has the largest percentage of open area.

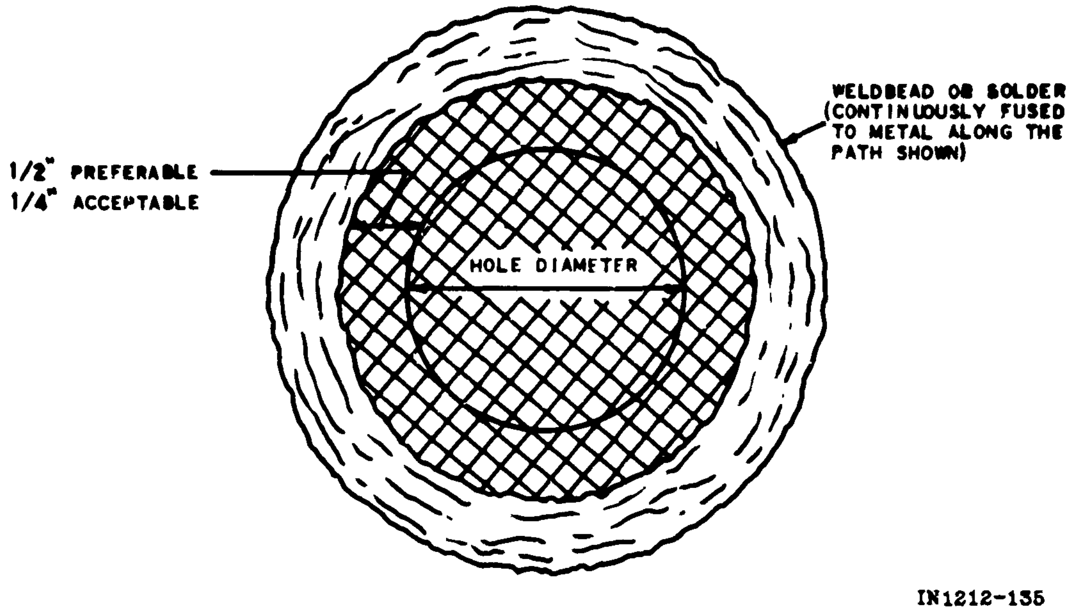



Figure 2-42. Typical Welded Screen Installation Over a Ventilation Aperture

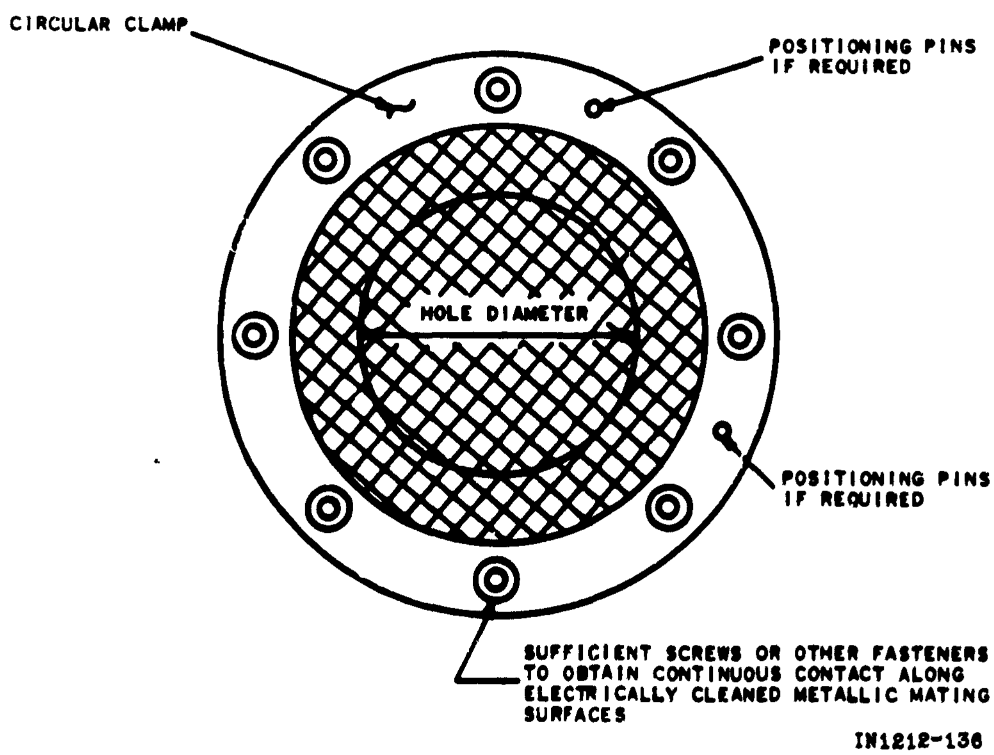



Figure 2-43. Typical Clamped Screen Installation Over a Ventilation Aperture

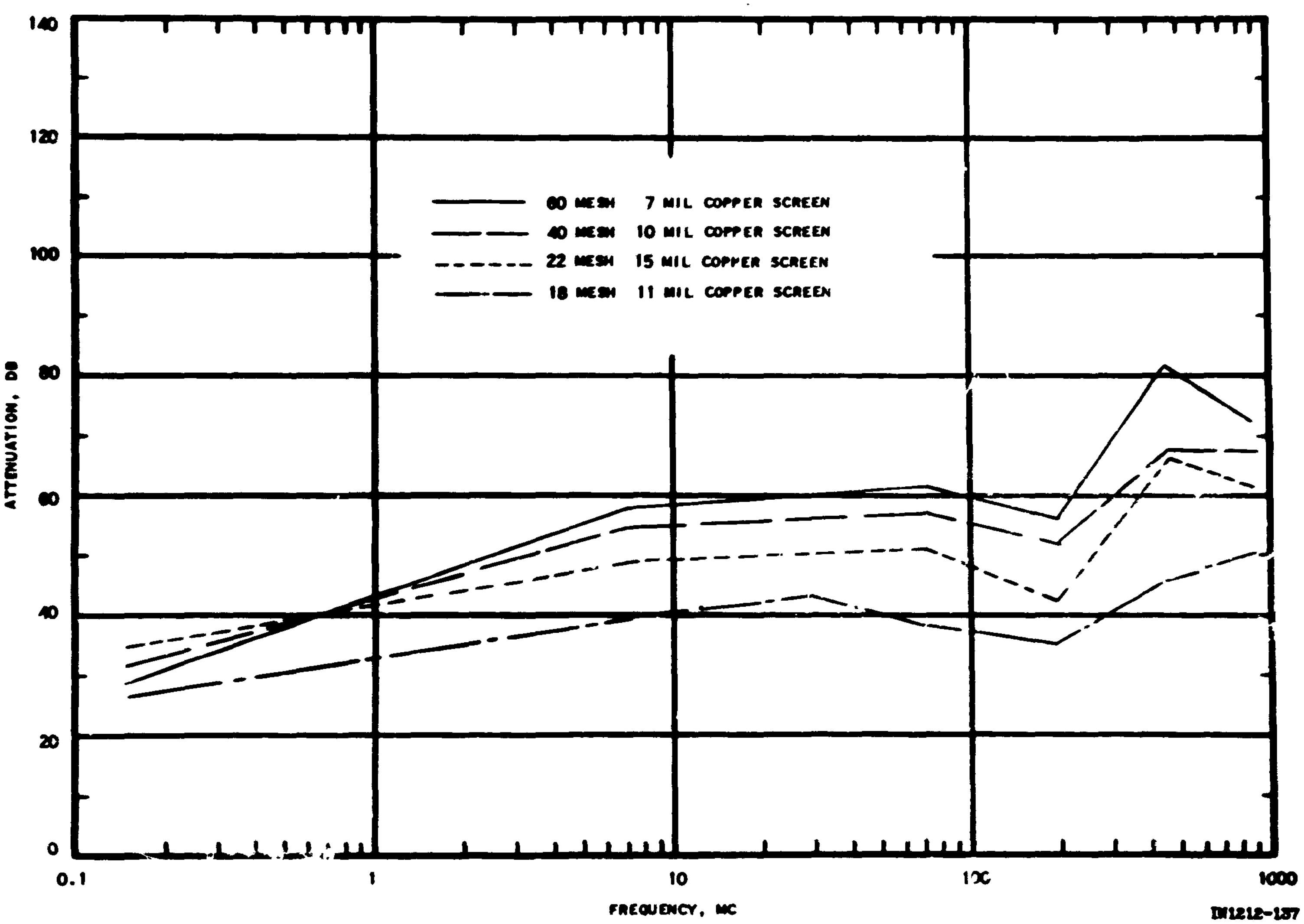



Figure 2-44. Shielding Effectiveness of Various Copper Screens

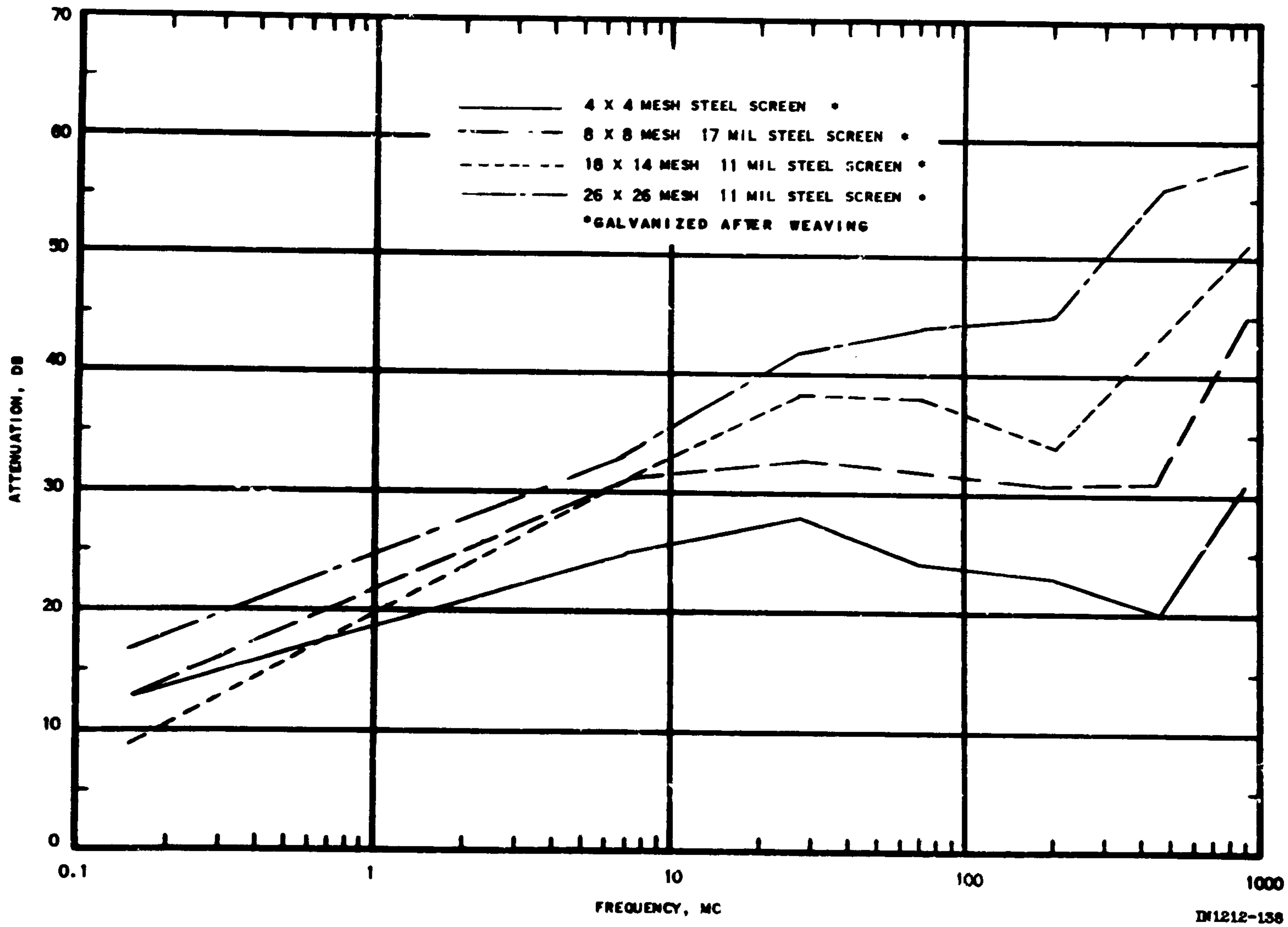



Figure 2-45. Shielding Effectiveness of Various Galvanized Steel Screens

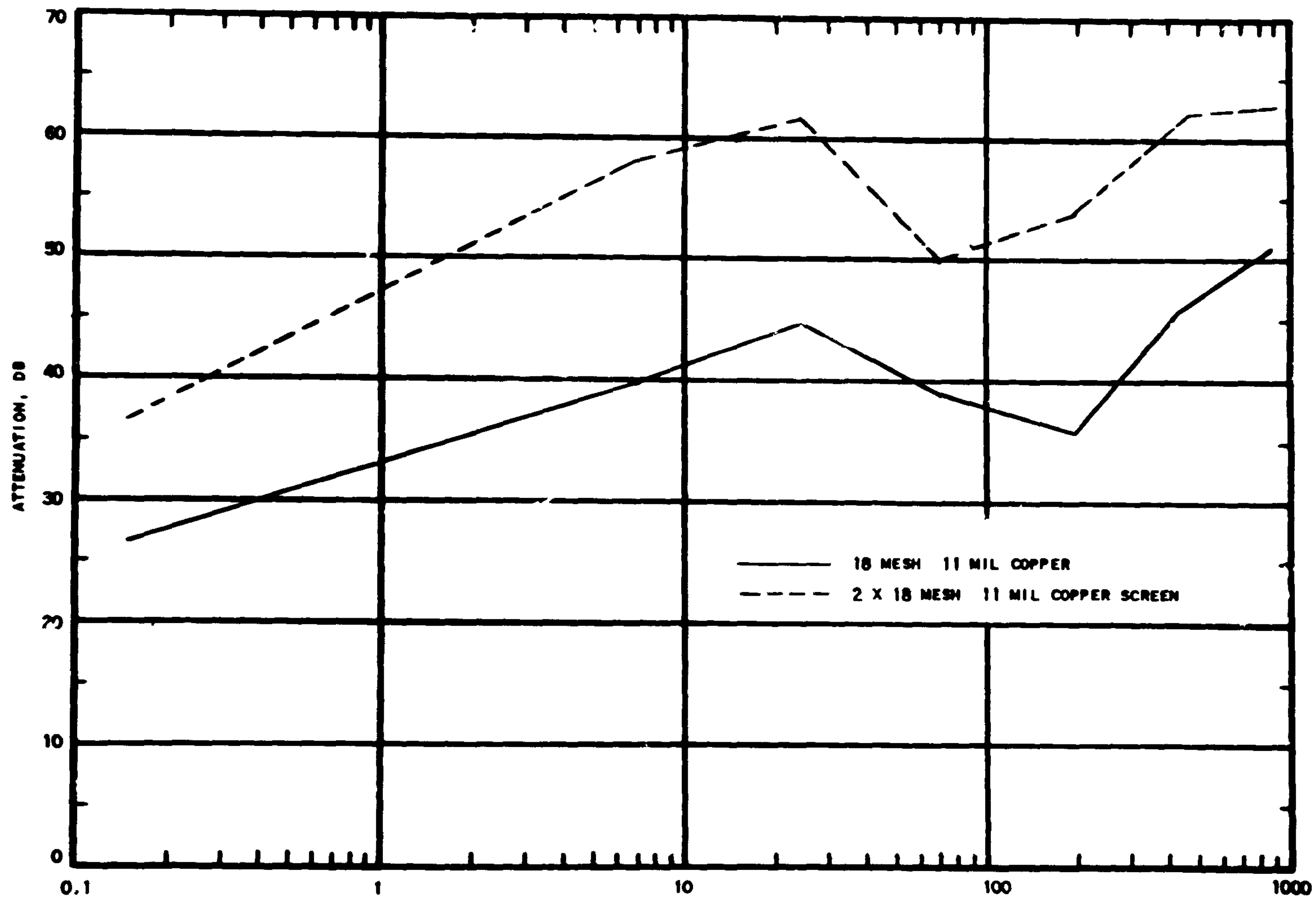



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Figure 2-46. Comparison of the Shielding Effectiveness of Single and Double 18 Mesh Copper Screening

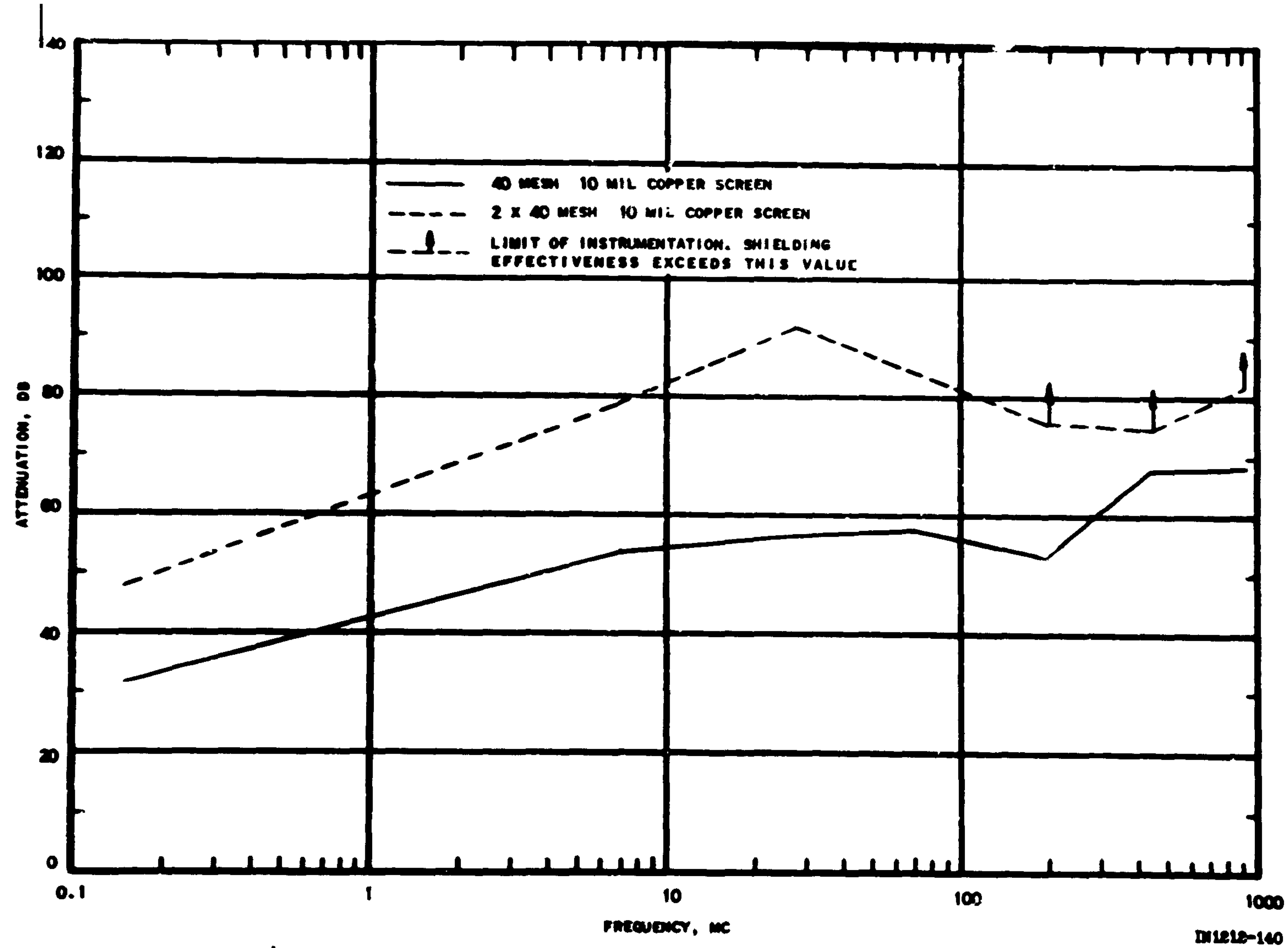



Figure 2-47. Comparison of the Shielding Effectiveness of Single and Double 40 Mesh Copper Screening

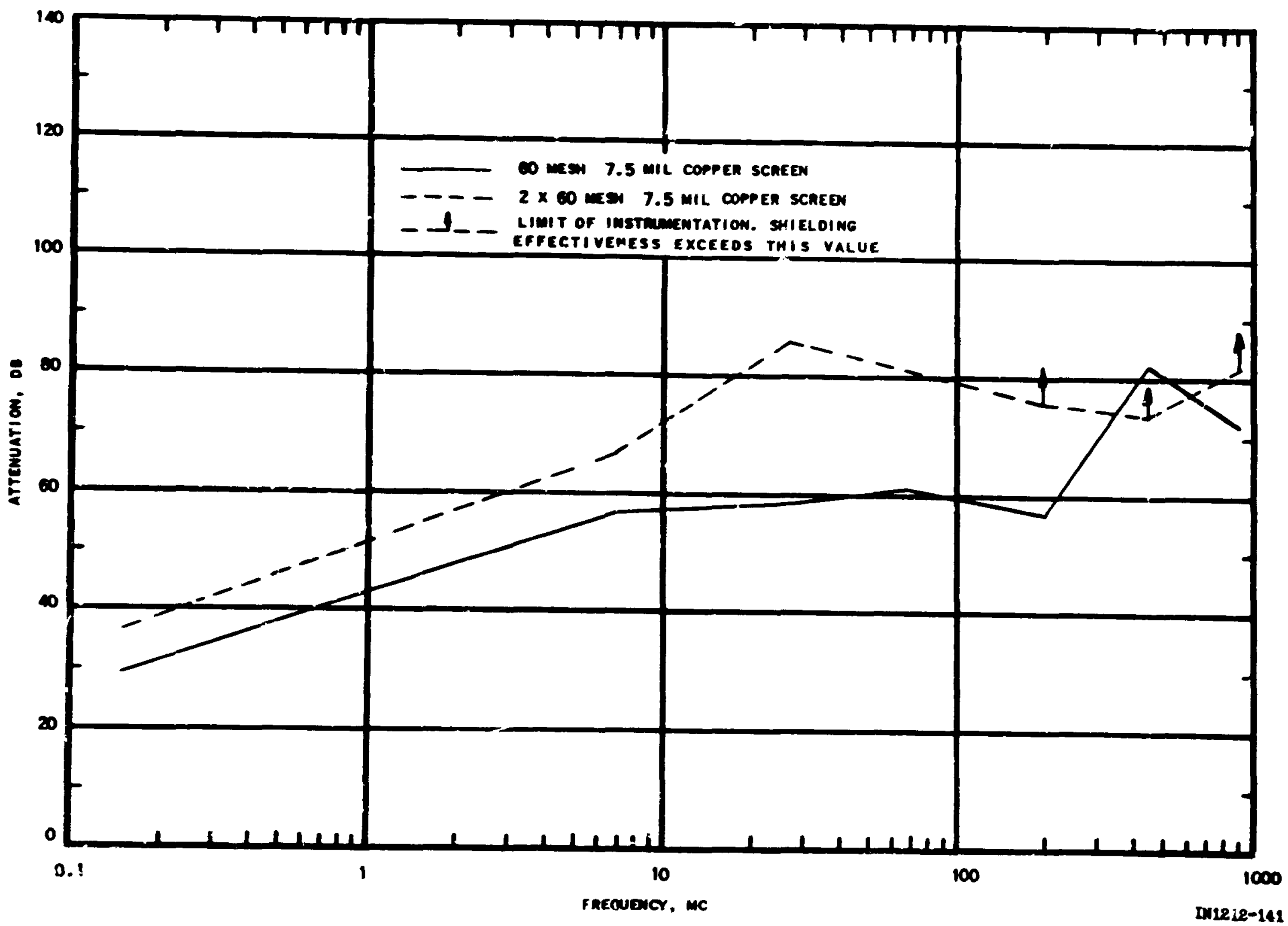



Figure 2-48. Comparison of the Shielding Effectiveness of Single and Double 69 Mesh Copper Screening

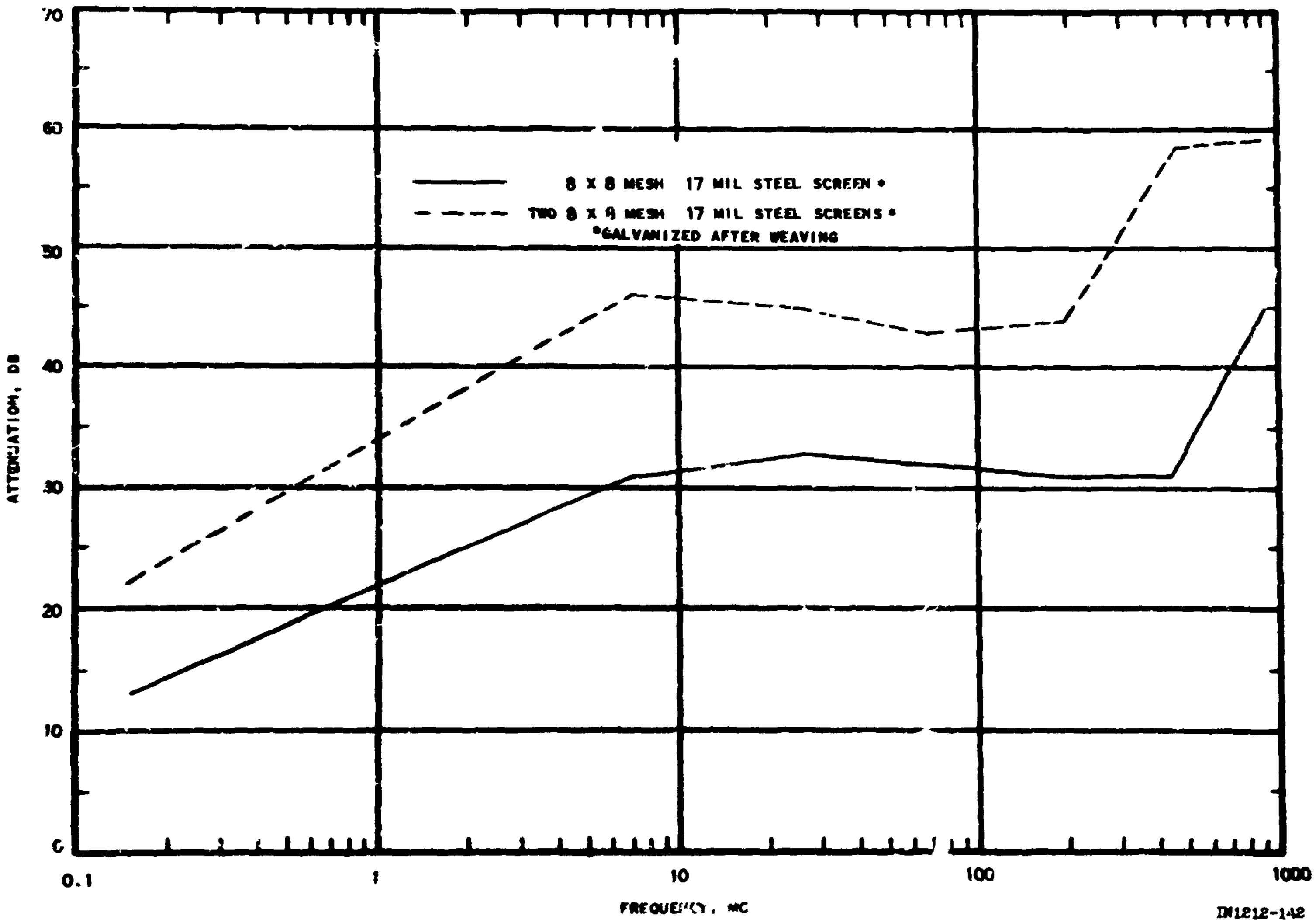



Figure 2-49. Comparison of the Shielding Effectiveness of Single and Double 8 Mesh Galvanized Steel Screening

TABLE 2-23. MESH, WIRE AND APERTURE SIZES FOR TYPICAL SCREENING MATERIALS

| Mesh | Wire Diameter (inches) | Size of Opening (inches) |
|---|---|---|
| 8 x 8 | 0.028 | 0.097 |
| 8 x 8 | 0.032 | 0.093 |
| 8 x 8 | 0.035 | 0.090 |
| 8 x 8 | 0.047 | 0.078 |
| 8 x 8 | 0.063 | 0.062 |
| 10 x 10 | 0.025 | 0.075 |
| 10 x 10 | 0.032 | 0.068 |
| 10 x 10 | 0.035 | 0.065 |
| 10 x 10 | 0.041 | 0.059 |
| 12 x 12 | 0.018 | 0.065 |
| 12 x 12 | 0.023 | 0.060 |
| 12 x 12 | 0.028 | 0.055 |
| 12 x 12 | 0.035 | 0.048 |
| 12 x 12 | 0.041 | 0.042 |
| 14 x 14 | 0.017 | 0.054 |
| 14 x 14 | 0.020 | 0.051 |
| 14 x 14 | 0.025 | 0.046 |
| 14 x 14 | 0.032 | 0.039 |
| 16 x 16 | 0.016 | 0.0465 |
| 16 x 16 | 0.018 | 0.0445 |
| 16 x 16 | 0.028 | 0.0345 |
| 18 x 18 | 0.017 | 0.0386 |
| 18 x 18 | 0.020 | 0.0356 |
| 18 x 18 | 0.025 | 0.0306 |
| 20 x 20 | 0.014 | 0.0360 |
| 20 x 20 | 0.016 | 0.0340 |
| 20 x 20 | 0.020 | 0.0300 |
| 22 x 22 | 0.015 | 0.0305 |

TABLE 2-24. MESH, WIRE AND APERTURE SIZES FOR SCREENING MATERIALS TESTED

| Metal | Mesh Size | Wire Diameter (mils) |
|---|---|---|
| Monel | No. 10 | 18 |
| Copper | No. 12 | 20 |
| Aluminum | No. 16 | 20 |
| Galvanized steel | 1/4-inch x 1/4-inch | 30 |
| Galvanized steel | 1/2-inch x 1/2-inch | 30 |
| Perforated steel[a] | 1/8-inch diameter holes | on 3/16-inch centers |
| Perforated aluminum[a] | 1/4-inch diameter holes | on 5/16-inch centers |
| Perforated aluminum[b] | 7/16-inch diameter holes | on 5/8-inch centers |
| Aluminum honeycomb[c] | 1/4-inch segregated cells | |

[a]60 mils thick

[b]37 mils thick

[c]1 inch thick

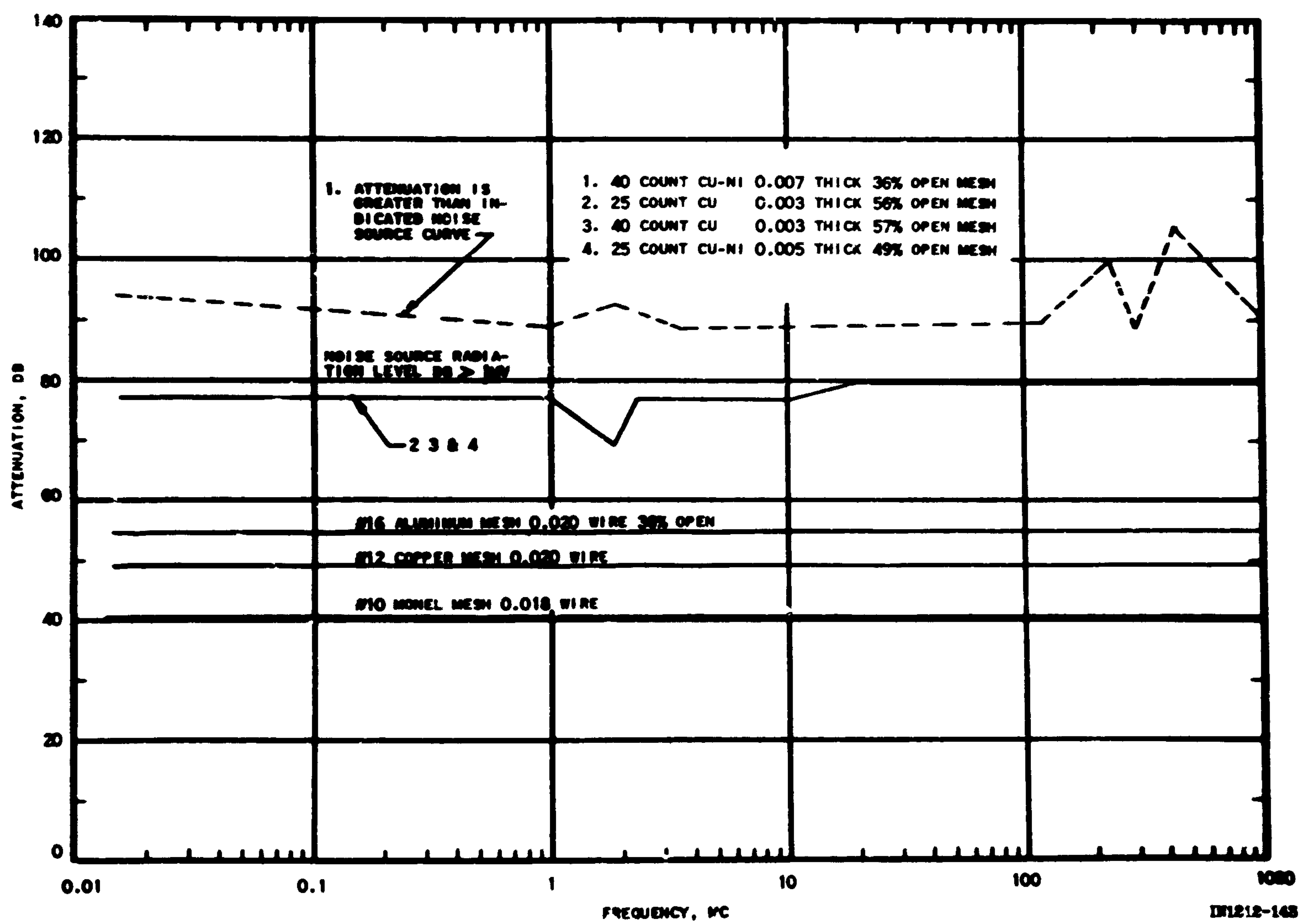



Figure 2-50. Attenuation Versus Frequency Curves For Various Screens

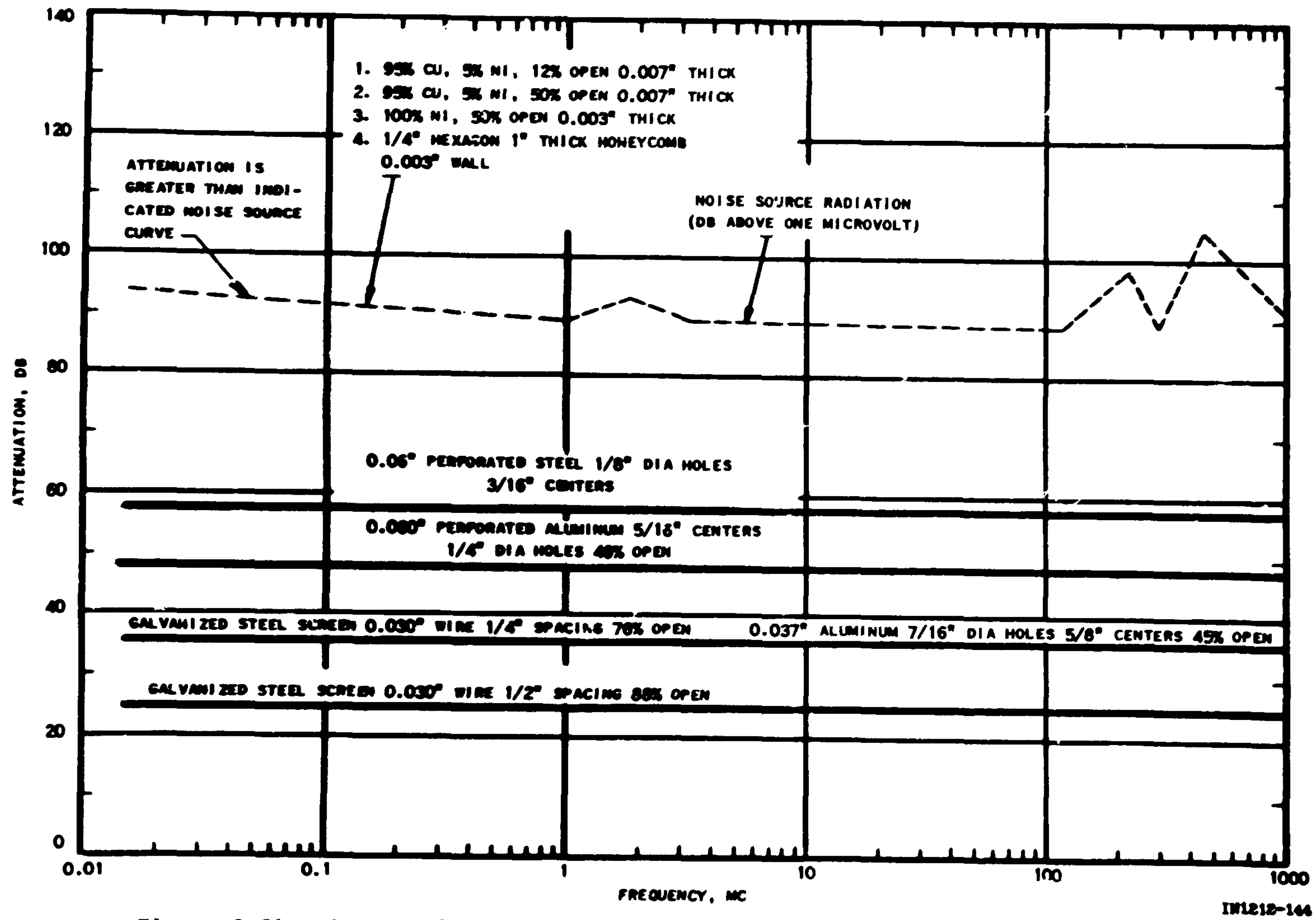



Figure 2-51. Attenuation Versus Frequency Curves for Various Screens and Honeycomb

(6) Figure 2-50 shows test results of materials. The 40-count copper-nickel 36 percent open sample has a range greater than 90 db. Such electroformed mesh materials are excellent as rf screening agents for panel meters, pilot lamps, counters, and other similar aperture discontinuities, or as simple inner-unit shields where small volumes of air transmission are necessary. The physical characteristics of typical perforated sheet stock are summarized in table 2-25.

(7) Figures 2-52 and 2-53 illustrate the measured magnetic field shielding effectiveness of screens of the various designs and metallic compositions listed in table 2-26 for the range of 0.15 to 1000 mc.

d. <u>Waveguide-Below-Cutoff Devices</u>.

(1) Electromagnetically, a small aperture is one which is smaller in its largest dimension than a signal wavelength. An aperture approaching or exceeding a wavelength in size must be covered by a fine mesh copper screen. Alternatively, a series of small unscreened apertures may be used instead of a single large hole; or waveguide attenuators may be used to shield large apertures. Waveguide attenuators can be designed to provide over 130 decibels of attenuation. The waveguide attenuator is also of considerable value when control shafts must pass through an enclosure. When an insulated control shaft passes through a waveguide attenuator, the control function can be accomplished with almost no interference leakage.

(2) In many cases, shielding screens introduce excessive air resistance, and sometimes greater shielding effectiveness may be needed than they can provide. In such cases, openings may be covered with specially designed ventilation panels (such as honeycomb) with openings that operate on the waveguide-below-cutoff principle. A sample panel is shown on figure 2-41. Honeycomb-type ventilation panels in place of screening:

TABLE 2-25. CHARACTERISTICS OF TYPICAL PERFORATED SHEET STOCK

| Hole Size (inches) | Hole Centers (inches) | Holes (square inch) | Open Area (%) |
|---|---|---|---|
| 0.014 | 0.034 | 890 | 13 |
| 0.014 | 0.042 | 665 | 10 |
| 0.018 | 0.045 | 504 | 13 |
| 0.018 | 0.055 | 380 | 10 |
| 0.020 | 0.034 | 890 | 27 |
| 0.020 | 0.040 | 650 | 20 |
| 0.020 | 0.042 | 665 | 20 |
| 0.020 | 0.045 | 504 | 15 |
| 0.020 | 0.049 | 500 | 15 |
| 0.020 | 0.055 | 380 | 12 |
| 0.020 | 0.085 | 144 | 4 |
| 0.020 | 0.089 | 148 | 4 |
| 0.023 | 0.040 | 650 | 26 |
| 0.023 | 0.049 | 500 | 20 |
| 0.024 | 0.037 | 710 | 32 |
| 0.024 | 0.040 | 650 | 28 |
| 0.024 | 0.045 | 504 | 22 |
| 0.024 | 0.046 | 542 | 24 |
| 0.024 | 0.049 | 500 | 21 |
| 0.024 | 0.055 | 380 | 17 |
| 0.024 | 0.059 | 331 | 14 |
| 0.024 | 0.063 | 292 | 13 |
| 0.024 | 0.079 | 160 | 8 |
| 0.024 | 0.079 | 186 | 8 |
| 0.024 | 0.085 | 144 | 6 |
| 0.024 | 0.088 | 148 | 7 |
| 0.024 | 0.110 | 82 | 4 |
| 0.024 | 0.110 | 94 | 4 |
| 0.025 | 0.045 | 504 | 28 |
| 0.025 | 0.055 | 380 | 20 |

TABLE 2-25. CHARACTERISTICS OF TYPICAL PERFORATED SHEET STOCK (cont'd)

| Hole Size (inches) | Hole Centers (inches) | Holes (square inch) | Open Area (%) |
|---|---|---|---|
| 0.025 | 0.059 | 331 | 17 |
| 0.025 | 0.085 | 144 | 7 |
| 0.025 | 0.088 | 148 | 8 |
| 0.028 | 0.048 | 420 | 26 |
| 0.028 | 0.059 | 331 | 20 |
| 0.028 | 0.102 | 94 | 6 |
| 0.028 | 0.102 | 110 | 7 |
| 0.030 | 0.059 | 331 | 23 |
| 0.030 | 0.102 | 94 | 7 |
| 0.030 | 0.102 | 110 | 8 |
| 0.032 | 0.071 | 465 | 35 |
| 0.032 | 0.059 | 331 | 26 |
| 0.032 | 0.102 | 110 | 9 |
| 0.033 | 0.059 | 331 | 27 |
| 0.034 | 0.059 | 331 | 29 |
| 0.035 | 0.059 | 331 | 31 |
| 0.036 | 0.055 | 330 | 37 |
| 0.036 | 0.055 | 380 | 37 |
| 0.036 | 0.059 | 331 | 33 |
| 0.036 | 0.069 | 245 | 24 |
| 0.036 | 0.099 | 119 | 12 |
| 0.036 | 0.119 | 70 | 7 |
| 0.036 | 0.119 | 82 | 8 |
| 0.037 | 0.079 | 186 | 22 |
| 0.039 | 0.069 | 245 | 30 |
| 0.039 | 0.074 | 213 | 26 |
| 0.039 | 0.079 | 186 | 23 |
| 0.039 | 0.099 | 119 | 15 |
| 0.039 | 0.108 | 98 | 12 |

TABLE 2-25. CHARACTERISTICS OF TYPICAL PERFORATED SHEET STOCK (cont'd)

| Hole Size (inches) | Hole Centers (inches) | Holes (square inch) | Open Area (%) |
|---|---|---|---|
| 0.039 | 0.130 | 59 | 7 |
| 0.039 | 0.130 | 68 | 8 |
| 0.039 | 0.237 | 20 | 3 |
| 0.041 | 0.079 | 186 | 25 |
| 0.043 | 0.079 | 186 | 28 |
| 0.043 | 0.091 | 140 | 21 |
| 0.045 | 0.079 | 186 | 30 |
| 0.045 | 0.091 | 136 | 22 |
| 0.048 | 0.079 | 186 | 33 |
| 0.048 | 0.091 | 140 | 25 |
| 0.048 | 0.079 | 186 | 36 |
| 0.050 | 0.102 | 110 | 21 |
| 0.050 | 0.119 | 82 | 16 |
| 0.050 | 0.138 | 60 | 12 |
| 0.050 | 0.177 | 36 | 7 |
| 0.050 | 0.201 | 28 | 6 |
| 0.052 | 0.079 | 186 | 38 |
| 0.052 | 0.091 | 140 | 23 |
| 0.052 | 0.102 | 110 | 23 |
| 0.052 | 0.119 | 82 | 17 |
| 0.052 | 0.138 | 60 | 13 |
| 0.052 | 0.157 | 46 | 10 |
| 0.053 | 0.091 | 140 | 30 |
| 0.053 | 0.110 | 94 | 21 |
| 0.055 | 0.091 | 140 | 34 |
| 0.055 | 0.110 | 94 | 23 |
| 0.055 | 0.126 | 64 | 15 |
| 0.055 | 0.157 | 46 | 11 |
| 0.055 | 0.189 | 32 | 8 |

TABLE 2-25. CHARACTERISTICS OF TYPICAL PERFORATED SHEET STOCK (cont'd)

| Hole Size (inches) | Hole Centers (inches) | Holes (square inch) | Open Area (%) |
|---|---|---|---|
| 0.057 | 0.099 | 119 | 31 |
| 0.059 | 0.092 | 136 | 37 |
| 0.059 | 0.099 | 119 | 33 |
| 0.059 | 0.102 | 110 | 30 |
| 0.059 | 0.119 | 82 | 23 |
| 0.059 | 0.157 | 46 | 13 |
| 0.059 | 0.170 | 35 | 10 |
| 0.059 | 0.170 | 40 | 11 |
| 0.059 | 0.205 | 27 | 8 |
| 0.061 | 0.102 | 111 | 30 |
| 0.063 | 0.102 | 110 | 34 |
| 0.063 | 0.119 | 82 | 26 |
| 0.063 | 0.126 | 72 | 23 |
| 0.063 | 0.177 | 36 | 11 |
| 0.063 | 0.205 | 26 | 9 |
| 0.063 | 0.217 | 24 | 8 |

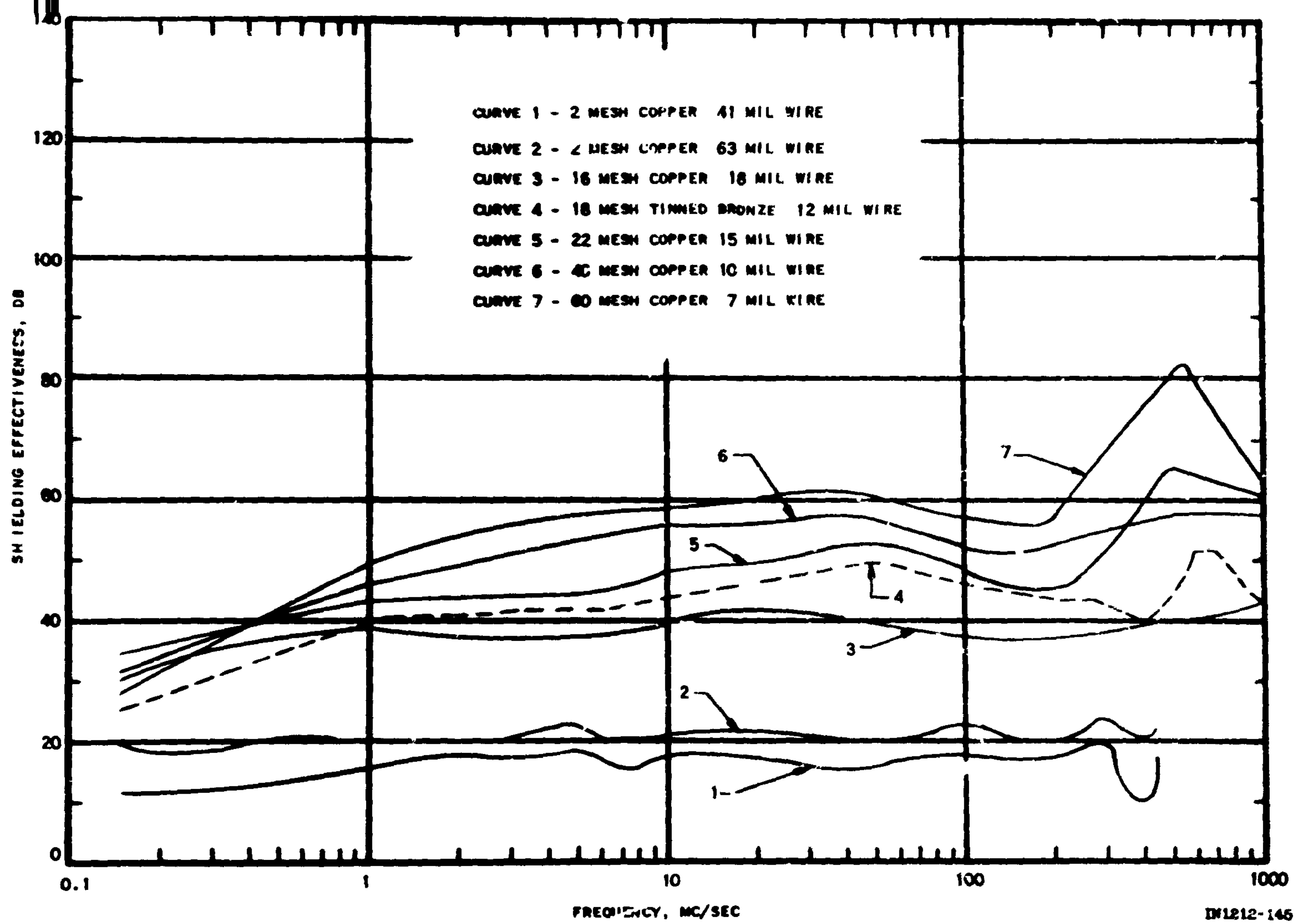



Figure 2-52. Minimum Shielding Effectiveness of Copper and Bronze Screening In a Magnetic Field

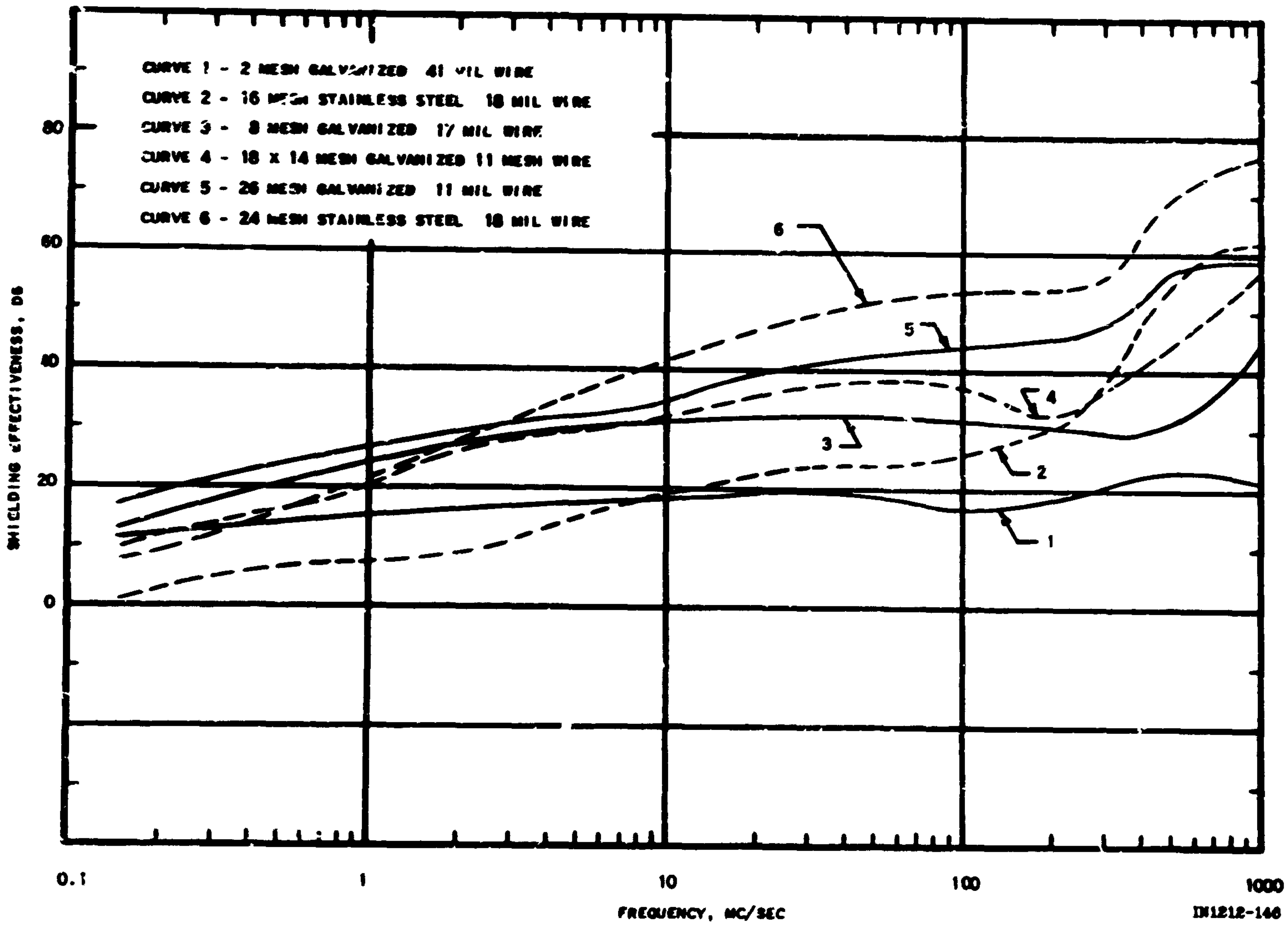



Figure 2-53. Minimum Shielding Effectiveness of Steel Screening in a Magnetic Field

TABLE 2-26. COPPER, BRONZE AND STEEL SCREENING USED IN MAGNETIC FIELD SHIELDING EFFECTIVENESS TEST MEASUREMENTS

| Metal | Mesh (wires/inch) | Wire Diameter (mils) |
|---|---|---|
| Copper | 2 | 41 |
| | 2 | 63 |
| | 16 | 18 |
| | 22 | 15 |
| | 40 | 10 |
| | 60 | 7 |
| Tinned Bronze | 18 | 12 |
| Galvanized Steel | 2 | 4 |
| | 8 | 17 |
| | 18 x 14 | 11 |
| | 26 | 11 |
| Stainless Steel | 16 | 18 |
| | 24 | 18 |

1) allow higher attenuation than can be obtained with mesh screening over a specified frequency range

2) allow more air to flow without pressure drop for the same diameter opening

3) cannot be damaged easily as can the mesh screen, and are therefore more reliable

4) less subject to deterioration by oxidation and exposure

All nonsolid shielding materials, such as perforated metal, fine mesh copper screening, and metal honeycomb, present an impedance to air flow. Metal honeycomb is the best of these materials because it enables very high electric field attenuations to be obtained through the microwave band with negligible drops in air pressure. Honeycomb, however, has the disadvantages of occupying far greater volume and costing far more than screening or perforated metal: panels of honeycomb vary in thickness from 1/4-inch to 2-3/8-inch, depending upon the attenuation desired. Also, it is often difficult to apply honeycomb paneling because flush mounting is required. Thus, screening and perforated sheet stock sometimes find application for purely physical design reasons, although honeycomb panels can achieve attenuations to 136 db, above 10 mc.

(3) Screened openings must usually be large to permit sufficient air to flow. When frequencies above 1000 mc are to be attenuated to a high degree, ventilation openings must be designed as waveguide attenuators operating below cutoff at their lowest propagating frequency. In this manner, shielding effectiveness of over 100 db can be obtained at frequencies of 10,000 mc: a 1/4-inch diameter tube, 1 inch in length, would have 102 db of shielding effectiveness at 10,000 mc; a 1/2-inch-diameter tube, 2-1/4 inches long, 100 db of shielding

effectiveness at 10,000 mc. Openings 1 inch or more in diameter would have little or no attenuation at 10,000 mc. To obtain an opening of sufficient size to admit the required volume of ventilating air, tubes should be placed side by side until sufficient air flow is achieved.

(4) The action of the waveguide operating below cut-off frequency (when it is used at a wavelength greater than its cutoff wavelength) is represented on figure 2-54 for a rectangular waveguide, and figure 2-55 for a circular waveguide, both for a $\frac{D}{W}$ ratio of 1. When the ratio is not 1 in a particular design problem, the value in decibels, obtained from the curve, must be multiplied by $\frac{D}{W}$ to arrive at the correct value of attenuation. The equation for a rectangular waveguide attenuator is:

$$A = 27.3 \frac{D}{W} \sqrt{1 - \left(\frac{Wf}{5910}\right)^2} \text{ decibels} \qquad (2\text{-}46)$$

where: D = depth of the waveguide in inches
W = largest inside cross-sectional dimension, in inches
f = frequency in megacycles

For a rectangular waveguide, to determine the depth required for 100 db attenuation at 1000 mc when the largest cross-sectional dimension is 0.125 inches:

$$A = 27.3 \frac{D}{W} \sqrt{1 - \left(\frac{Wf}{5910}\right)^2} \text{ decibels}$$

$$\frac{A}{D} = \frac{27.3}{W} \sqrt{1 - \left(\frac{Wf}{5910}\right)^2} \quad \frac{\text{decibels}}{\text{inch}}$$

$$= \frac{27.3}{(.125)} \sqrt{1 - \left[\frac{(.125)\ (1000)}{5910}\right]^2}$$

$$= 210 \ \frac{\text{db}}{\text{inch}} \qquad (2\text{-}47)$$

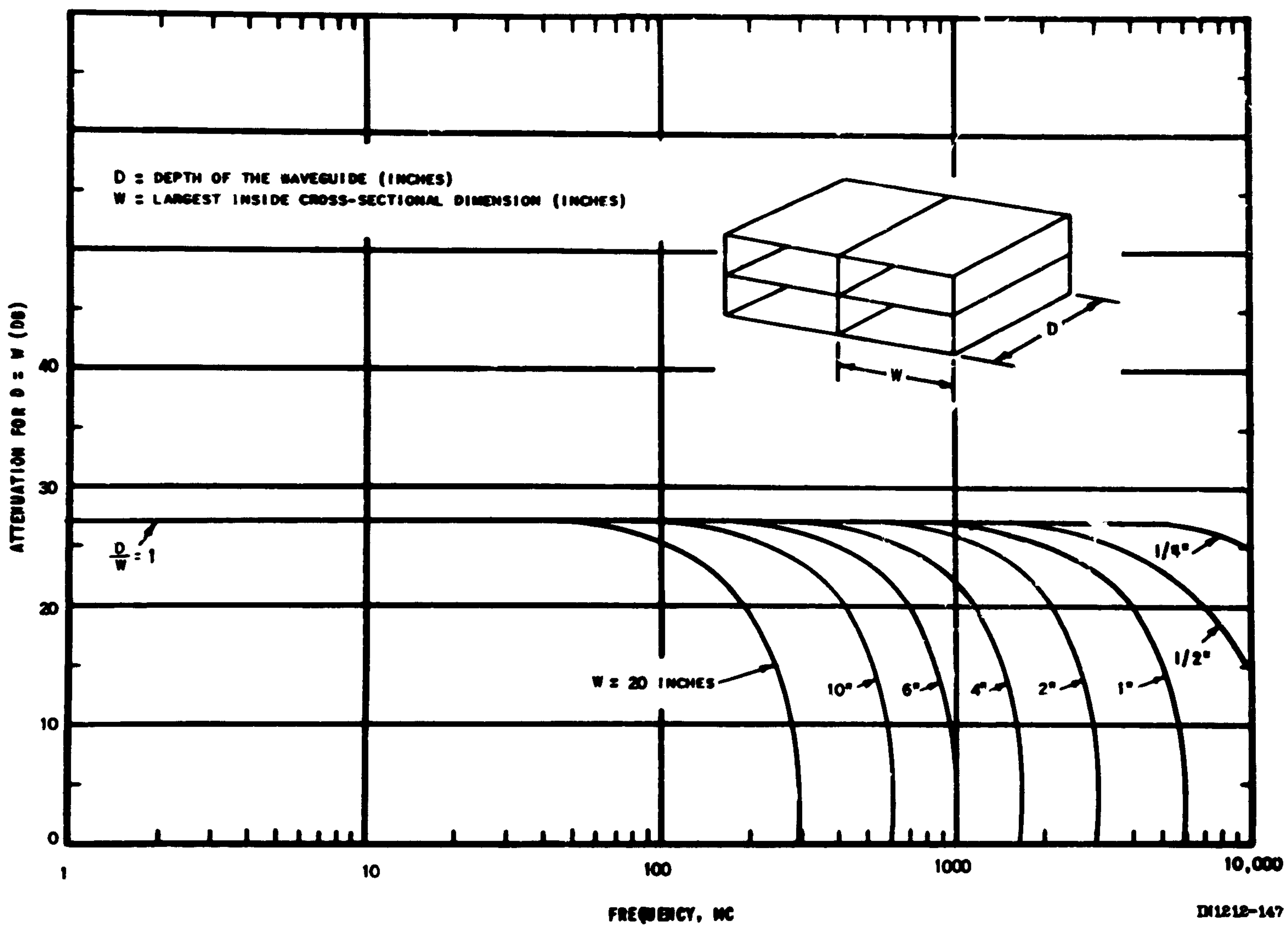



Figure 2-54. Attenuation-Rectangular Waveguide

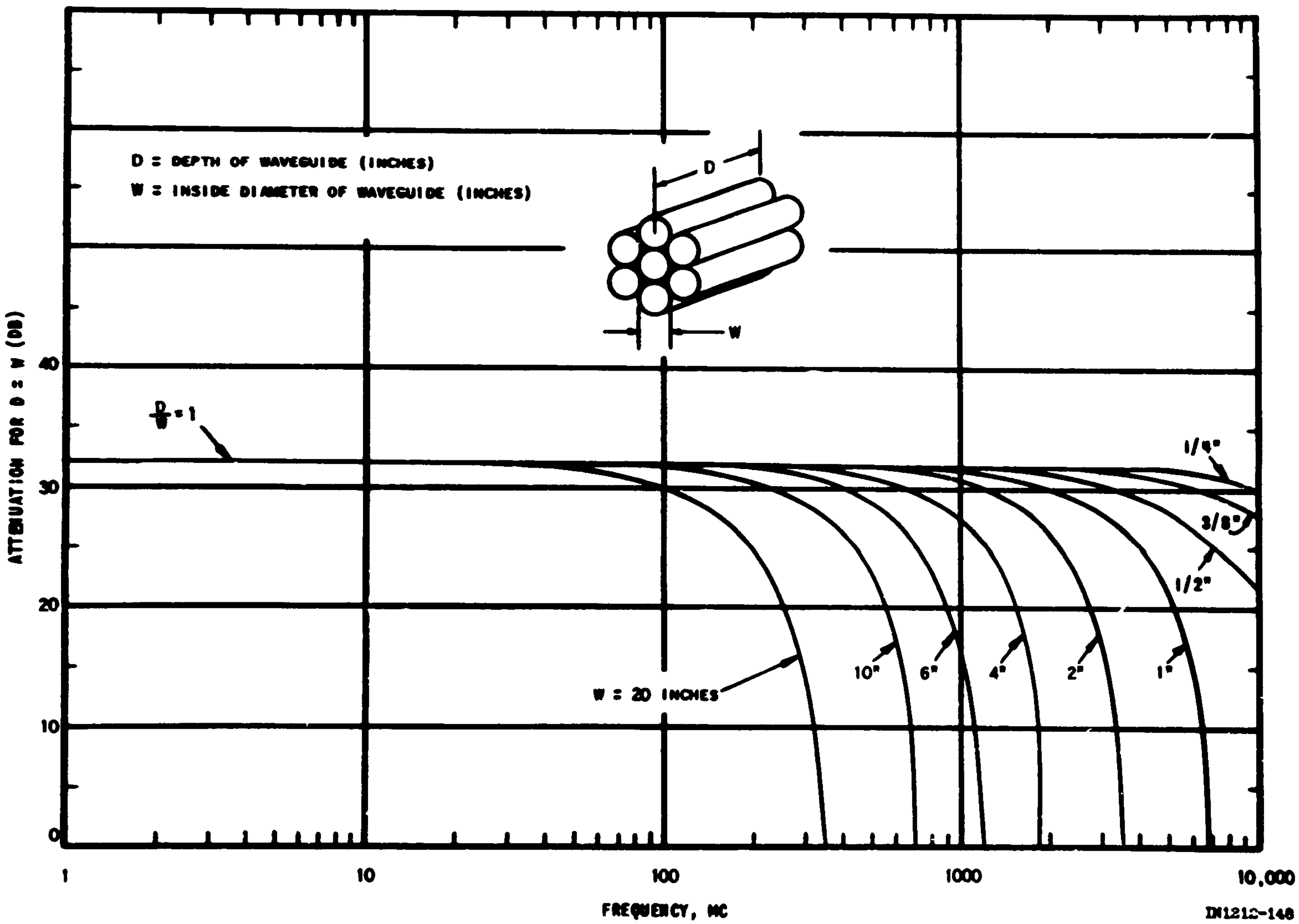



Figure 2-55. Attenuation-Circular Waveguide

To obtain 100 db attenuation:

$$\frac{100 \text{ db}}{D} = 210 \frac{\text{db}}{\text{inch}} \tag{2-48}$$

$$D = \frac{1 \text{ inch}}{210 \text{ db}} \; ; \; 100 \text{ db} \approx \frac{1}{2} \text{ inch}$$

The equation for a circular waveguide attenuator is:

$$A = 31.95 \frac{D}{W} \sqrt{1 - \left(\frac{Wf}{6920}\right)^2} \quad \text{decibels} \tag{2-49}$$

where: D = depth of the waveguide in inches

W = inside diameter of waveguide in inches

f = frequency in megacycles

These waveguides function as high-pass filters. All frequencies below the cutoff frequency are attenuated. The lowest cutoff frequency in megacycles for rectangular and circular waveguides are respectively:

$$f_c = \frac{5910}{\text{longest dimension of rectangle}} \tag{2-50}$$

and

$$f_c = \frac{6920}{\text{diameter of circular opening}} \tag{2-51}$$

where $f_c$ is in mc, and the dimensions are in inches.

The maximum operating frequency should be 1/10 of the cutoff frequency. Although the attenuation is near maximum at 1/3 of cutoff frequency, the 1/10 value is advisable to provide a safety factor. In the waveguide, the attenuation-per-unit length for an operating frequency, f, (in mc), below the cutoff frequency of the waveguide is:

$$\alpha = \text{db/inch of sleeve length} = 0.00463 \, f \sqrt{\left(\frac{f_c}{f}\right)^2 - 1} \tag{2-52}$$

(5) Figures 2-54 and 2-55 indicate the frequency range over which any particular opening size may be useful. It is assumed that the conductivity and electrical thickness of the metal walls between openings is sufficient at frequencies as low as 1 mc. The flat characteristic of the curves continues to the lowest frequency at which this condition is met. This waveguide-below-cutoff approach is recommended as a more reliable method of shielding, with higher attenuation and minimum air pressure drop. One manufacturer has reported a maximum head loss of 0.1 inch of water at 1200 feet/minute air velocity. Control shafts, that protrude through a hole in a shielded equipment enclosure, can be of metal, grounded by metallic fingers, or be of nylon, teflon, or other dielectrics inserted through the metallic tube or shield waveguide (figs. 2-56 and 2-57). The waveguide approach may be considered where holes must be drilled in an enclosure. If the metal thickness is sufficient to provide a tunnel with adequate length, it will effectively serve as a waveguide attenuator. For example, a metal wall 3/16-inch thick would permit a 1/16-inch diameter hole to be used without excessive leakage. This approach should be considered where it is necessary to confine extremely intense interference sources.

## 2-21. Fasteners

a. To provide adequate contact between mating metallic surfaces, it is recommended that sufficient fasteners be used to ensure good contact and, at the same time, care must be taken to avoid buckling between the surfaces. The spacing between fasteners must be much closer as the materials become thinner because of bending under pressure (fig. 2-58). The spacing of the fasteners should also preclude the possibility of radiation when the metal surfaces spread apart as a result of vibration or misalignment.

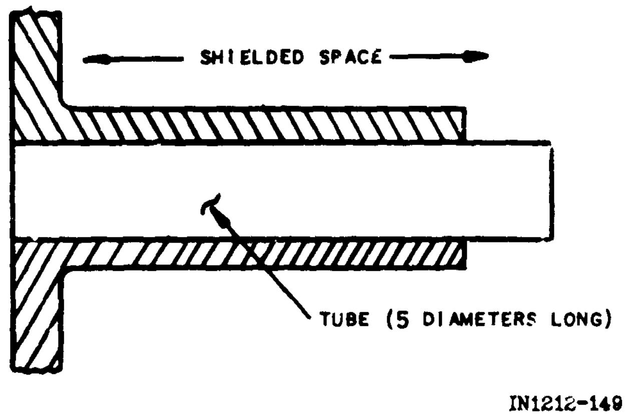



Figure 2-56. Tube Acting as Waveguide Attenuator

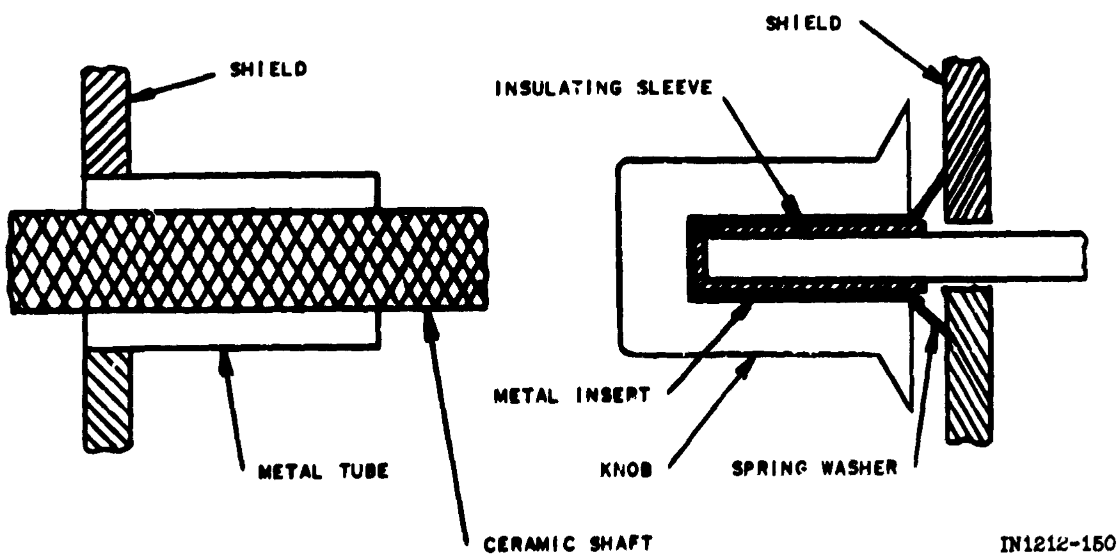



Figure 2-57. Shaft Feedthrough Techniques

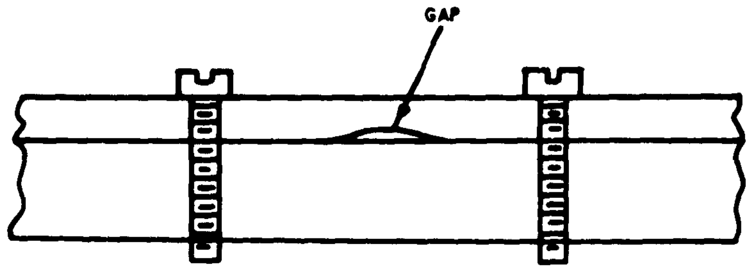
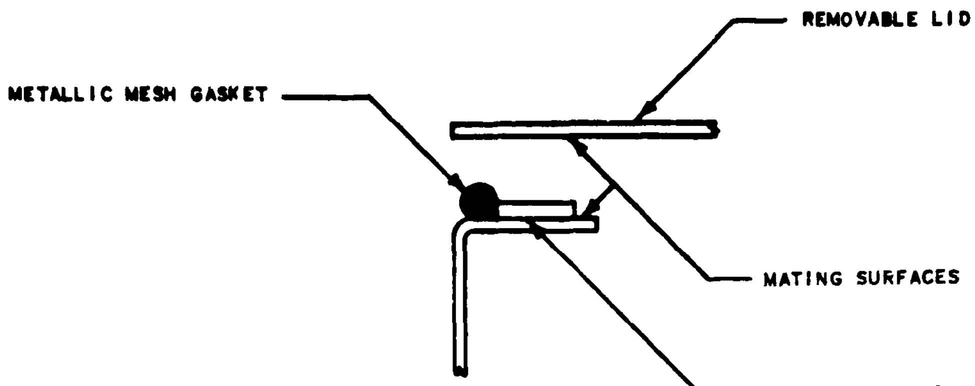
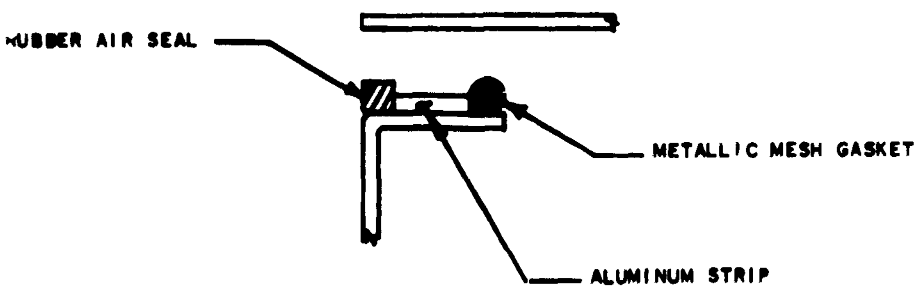



Figure 2-58. Buckling Between Fasteners With Thin Material

b. When it is necessary to join several parts of a complete shield, the first consideration should be to minimize the number of joints. When joints are made, the most important requirement is that a continuous metal-to-metal contact be maintained along a continuous line. When the pressure is maintained by screws or bolts, a sufficient number must be used to ensure high unit pressure even at the points farthest away from any screw or bolt. Lack of stiffness of mating members produces distortion of mating surfaces, which results in bulging and insufficient pressure for preserving good electrical contact. The design of these joints can be simplified by employing conductive gaskets (electronic weather stripping), especially where close fabrication tolerances need not be maintained. Such gaskets include metallic textile gaskets and knitted wire mesh, which are available in many different materials such as copper, monel, silver-plated brass, and beryllium copper. These gaskets can be combined with or imbedded in rubber or plastic to serve as water, air and oil seals, as well as impenetrable interference shields. In addition to shielding against interference, it is often necessary to provide an air seal at equipment seams. A method of achieving this is shown on figure 2-59. Numerous combinations of finger and mesh gasket designs are commercially available so that any sealing problem can be successfully resolved with proper design of mating surfaces.




A. MESH GASKET APPLICATION.




B. COMBINATION GASKET FOR AIR PRESSURE AND RFI SEAL.

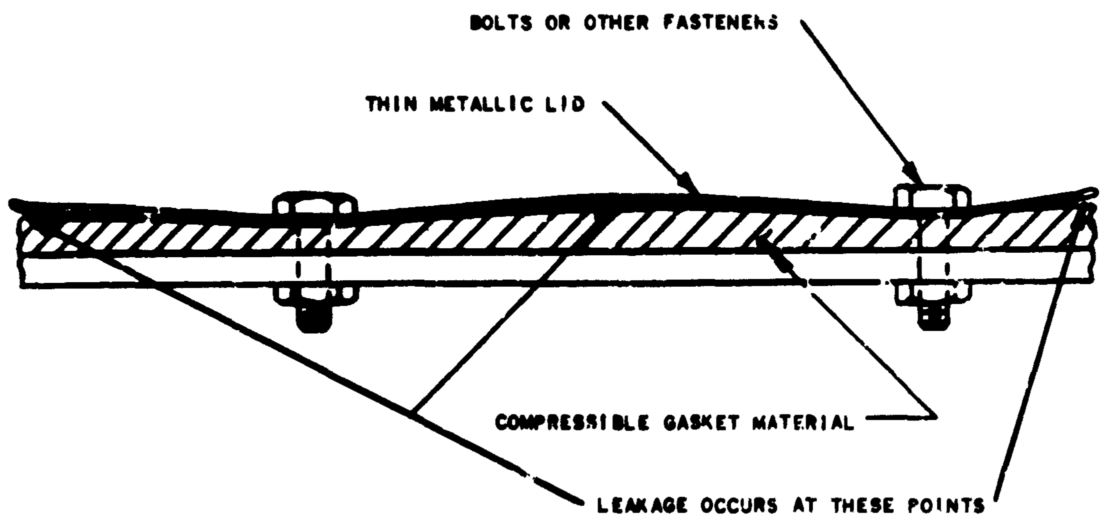
IN1212-151

Figure 2-59. Seam Treatment With Mesh Gaskets

c. Successful interference leakage reduction by the use of gasketing material is dependent on the achievement of good metallic contact along the entire length of the seam. Parameters governing the quality of the final design are shown on figure 2-60. In some cases, the general requirement for fasteners approximately every 4 inches along the seam can be reduced provided a good metallic seal exists between the metal surfaces and the conductive gasket. It is unwise to rely solely on screws, bolts, nuts, and internal or external toothed lockwashers for electrical bonding because they are subject to mechanical vibration and chemical deterioration. It is difficult to maintain a uniform pressure distribution over a large surface contact area unless hold-down screws are very closely spaced (fig. 2-60). Even when the screws are closely spaced, there is always a higher pressure in the areas immediately under the screws and a tendency of the cover to bow in the spaces between the screws. It is almost impossible to make a joint of this nature airtight unless the covers are sufficiently rigid to preclude bowing between screws and a thin film of oil is placed between the surfaces. Such an oil film will fill machine marks and indentations, or scratches on metal surfaces, and will make a good pressure seal; it will also, however, electrically insulate one section from another unless a special electrically conductive type of oil is used. To prevent electrolytic action, the two sections should be of the same metal, otherwise the parts will corrode.

2-22. Conductive Gaskets

a. General. In the past, when a pressure seal was desired, a compressible gasket was used between the two surfaces. The compressible gasket, being of neoprene or similar material, insulated those portions of the joint between bolt holes from each other, rendering the shielding ineffective. There are two types of compressible gaskets: the flat gasket and the groove gasket. The groove gasket is installed as shown on figure 2-62. Joints with insulating groove gaskets are much better than ones with flat, insulating gaskets as there is a reasonable possibility of metal surfaces making contact between some of the bolt holes. This possibility is considerably increased if the volume of the groove is greater than the volume of the gasket, and the gasket molding process is accurately controlled so that it will have uniform cross-sectional area, preventing the gasket from bulging out over the edges of the groove and insulating one metal surface of the shielding assembly from the other. Because practically all electronic equipment must be designed with seams or openings to facilitate inspection, cooling, data output, metering, tuning, or other functions, and because these seams or openings usually represent the weakest points in the overall shield design, it is imperative that they be shielded to prevent entry or exit of interfering, or potentially interfering, electromagnetic energy. If shielding is not accomplished, other interference control measures may prove unsuccessful. In sealing such items as seams and panel joints to interference, conductive gaskets are almost always required to ensure continuous low-impedance contact between mating metal surfaces.




NOTE: VIEW PURPOSELY EXAGGERATED TO DEMONSTRATE IMPERFECT SEAL CONDITIONS

IN1212-152

Figure 2-60. Improper Gasket Application

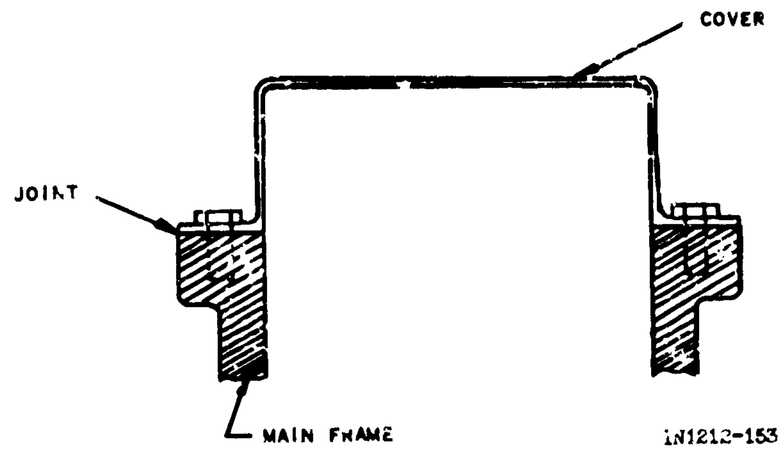



Figure 2-61. Flat-Flange Type Joint

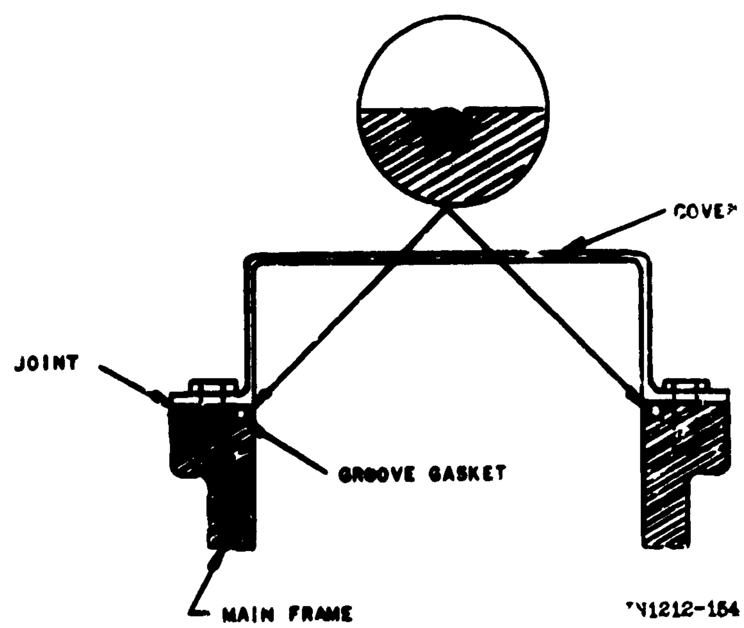



Figure 2-62. Flat-Flange Type Joint With Groove Gasket

b. Conductive Gasket Applications. In an attempt to improve shielding characteristics of flange type joints, such as the one presented on figure 2-62, a flat conducting gasket, consisting of an aluminum screen impregnated with neoprene was developed. It was found that the mesh was so fine that the neoprene was restricted in movement; and the gasket had little resilience and was a poor pressure seal. Later adaptations of this gasket were made of 16-mesh aluminum screen impregnated with neoprene; these were satisfactory.

(1) For satisfactory assembly, a gasket should be retained in a slot or groove. Figure 2-63 illustrates arrangements for use of this type of gasket in a flange and cover-plate joint. Typical closures using resilient conductive electronic weatherstrips are depicted on figure 2-64. In designing gaskets, it is necessary to provide minimum gasket thickness to allow for expected surface discontinuities of the joint; to provide correct height and pressure; and to allow for frequency of use. The gasket materials selected should be corrosion resistant, conductive, and possess an adequate degree of strength, resiliency, and hardness. Gasket characteristics may be classified on the basis of frequency and manner that the joint will be expected to be opened and closed during its lifetime. Gaskets may be held in place by sidewall friction, an attachment fin, or positioned by a shoulder as shown on figure 2-65. Gaskets are available in various configurations, including those shown on figure 2-66.

(2) Resilient metallic gaskets provide a simple, efficient, and inexpensive method for sealing joints in a shield, especially when the gasket is designed into the shield and not added as an afterthought. A rule of thumb that applies specifically to fluid gaskets can also be applied to conductive gaskets: the greater the compressibility, the greater the sealability.

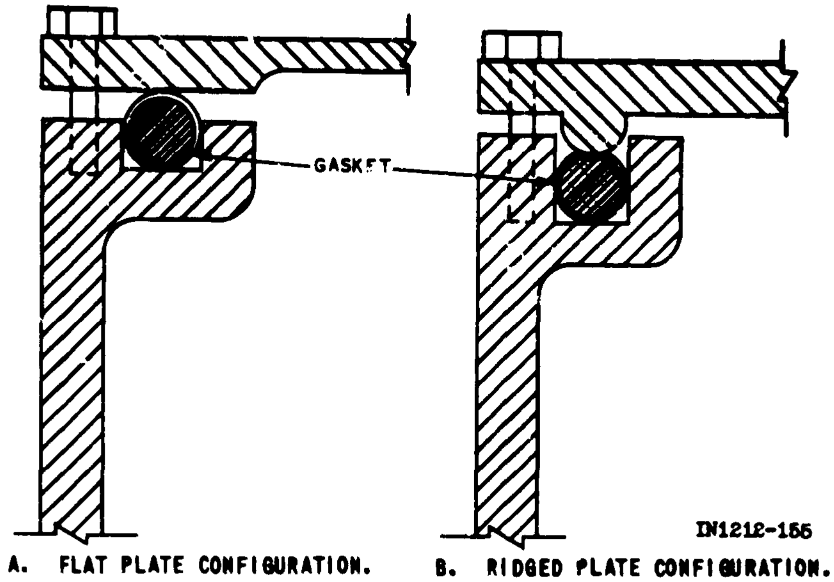



Figure 2-63. Typical Groove Type Gasket Applications

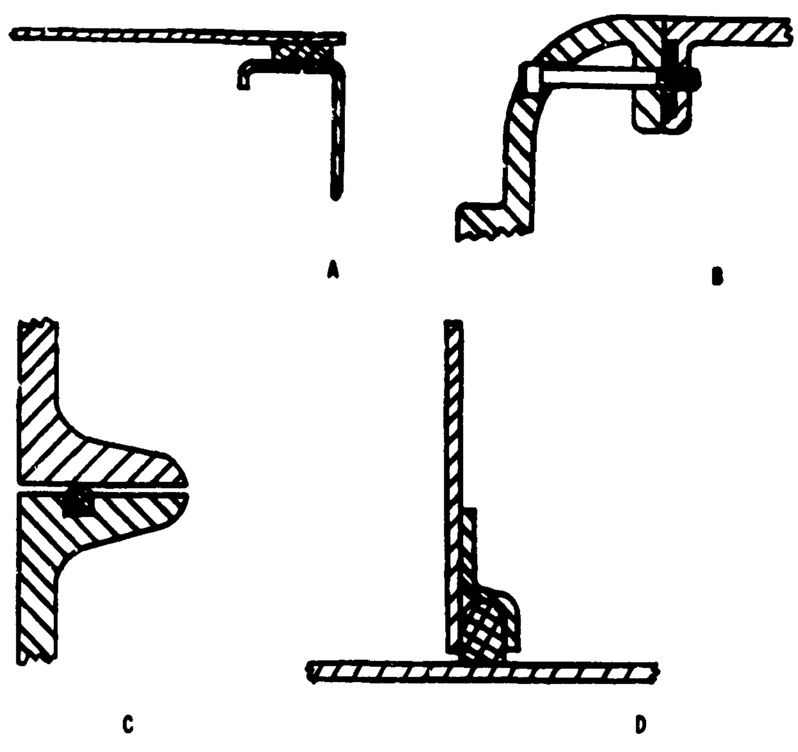
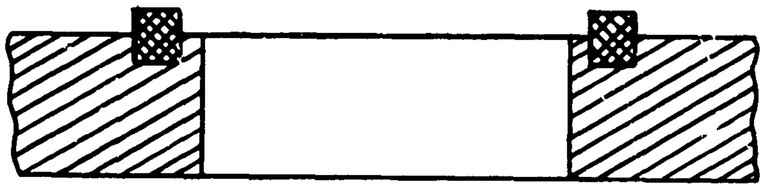
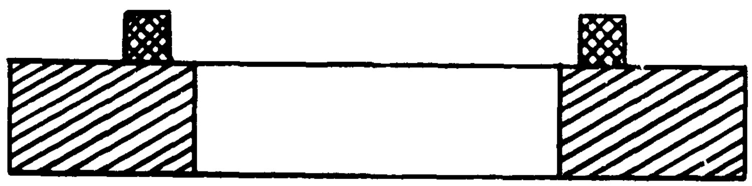


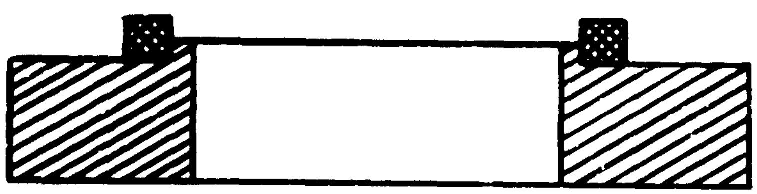

Figure 2-64. Typical Conductive Gasket Applications



A. HELD IN PLACE IN SLOT BY SIDE-WALL FRICTION



B. HELD IN PLACE BY SOLDERING (AMOUNT OF SOLDER USED MUST BE CAREFULLY CONTROLLED TO PREVENT ITS SOAKING INTO THE GASKET)



C. POSITIONED BY SHOULDER IN1212-157

Figure 2-65. Typical Gasket Mounting Methods

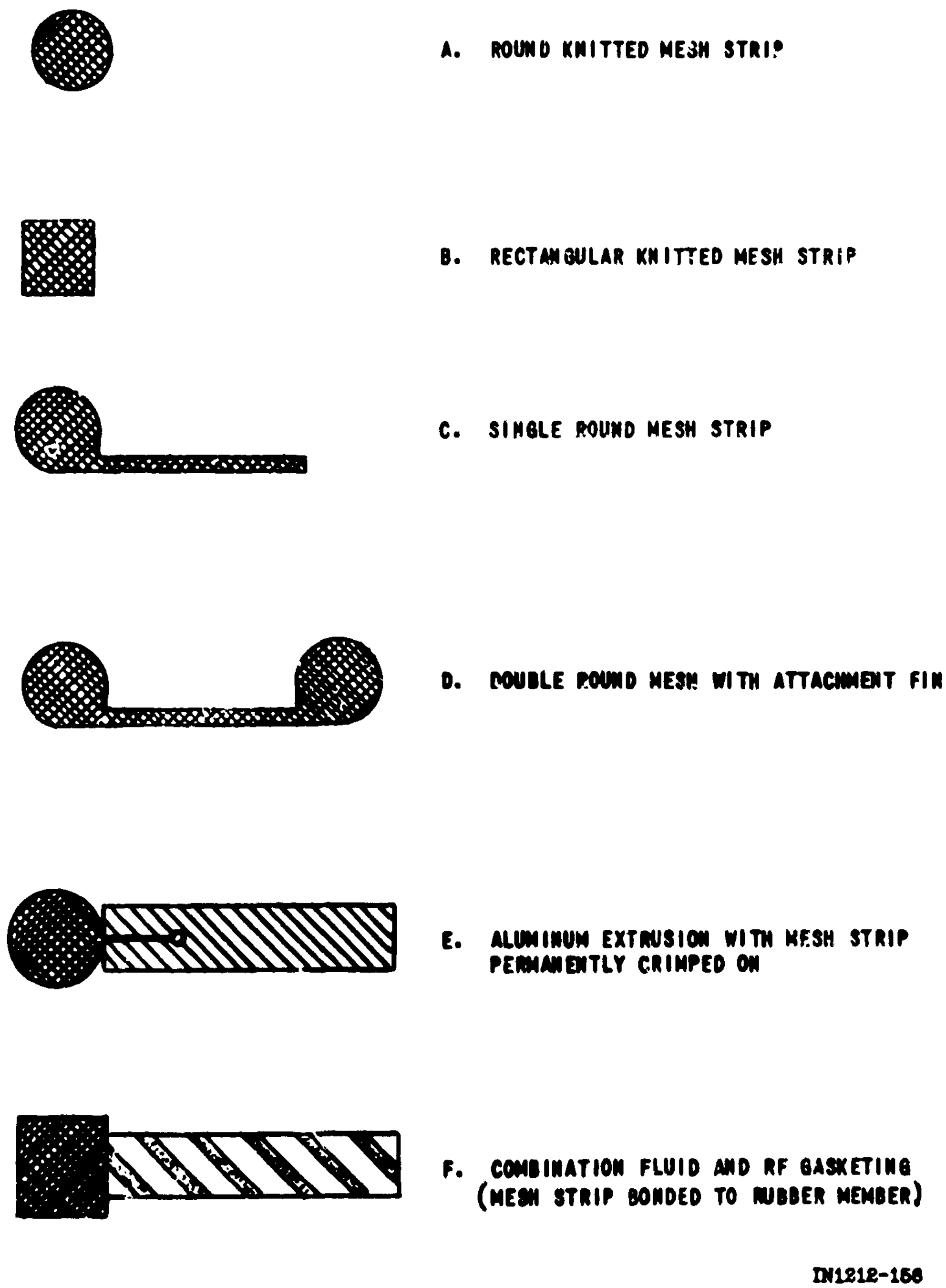



Figure 2-66. Typical Electrical Weatherstripping Applications

This principle is illustrated by figure 2-67, which depicts a simulated joint and three gaskets. Gasket A is one-half the height of B and C and is very resilient; gasket C has the same resiliency as A; and gasket B is harder than A. For simplicity, assume gaskets A and C have twice the resiliency of B. They compress 50 percent under the force F applied to the joint, while B compresses only 25 percent down to 75 percent of original height. Figure 2-68 shows gasket A compressed to 50 percent at the point of maximum compression; this is not sufficient to seal the joint fully. Gasket B is then inserted (fig. 2-69), but, because it compresses only 25 percent under the same force, its greater height does not result in more sealability (0.5h = 0.25 x 2h). In figure 2-70, gasket C is compressed 50 percent, the same percentage as A, because they are equally resilient. Because C is twice as thick as A, the same percentage of compression results in twice the actual compression . In the example (fig. 2-70), this is sufficient to effect a seal. Figures 2-67 through 2-70 illustrate the basic axiom of gasket design: the gasket must be compressible enough to conform to the irregularities of both surfaces under the applied force. This axiom can be applied to all conductive gaskets with the additional requirement that contact pressure be high enough to make adequate contact, even in the presence of nonconducting corrosion films.

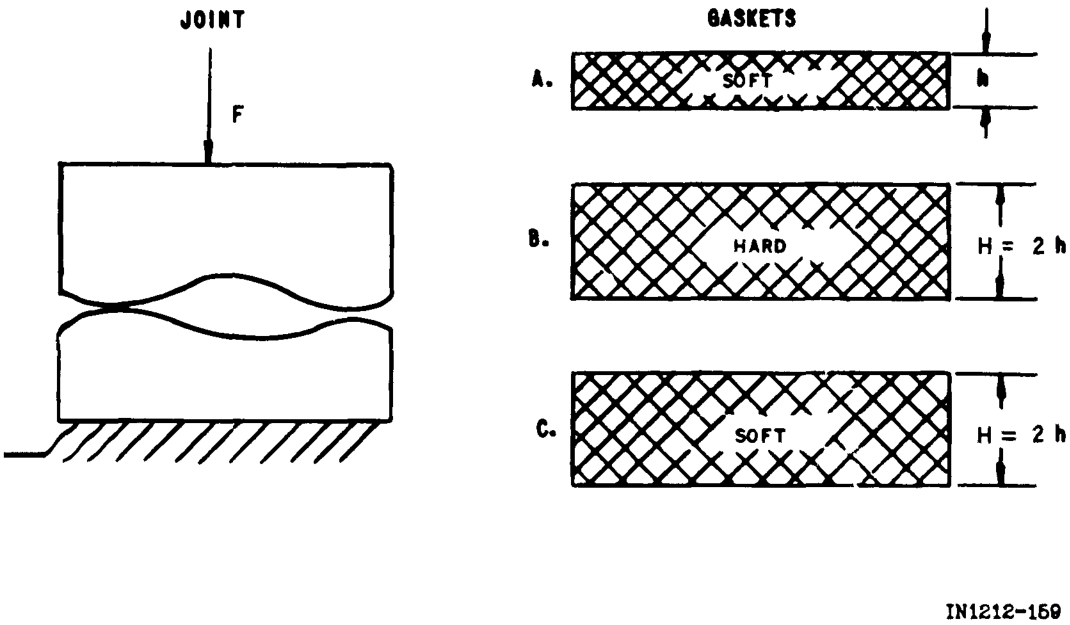



Figure 2-67. Gasket Compressibility Versus Sealability

(3) Although soldering gaskets in place is seldom recommended, it is included here because the method is often suggested. If the amount of solder and length of time it will be molten can be controlled accurately, soldering might be worthwhile. Care must be taken to prevent solder from soaking into the gasket and destroying its resiliency.

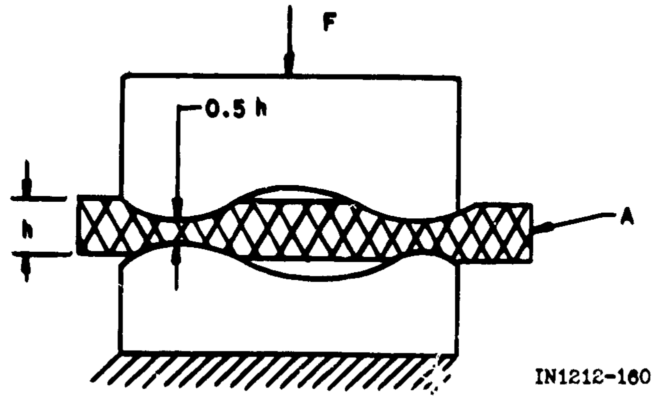



Figure 2-68. Gasket A in Simulated Joint

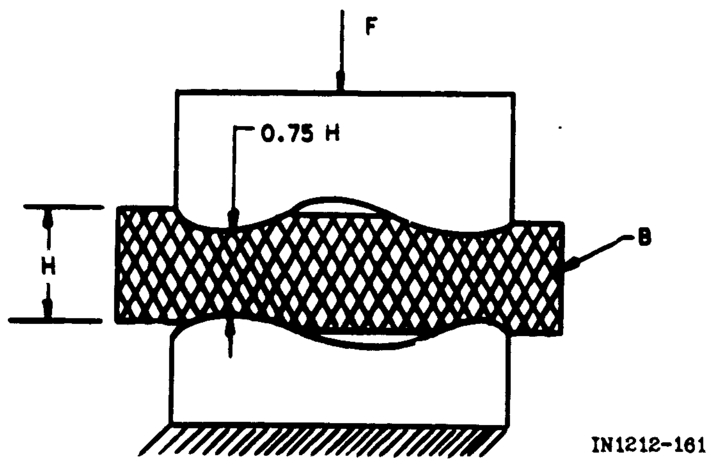



Figure 2-69. Gasket B in Simulated Joint

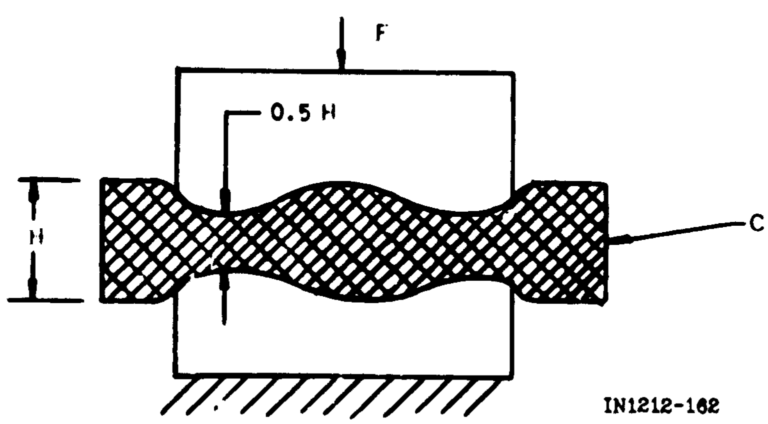



Figure 2-70. Gasket C in Simulated Joint

(4) Figure 2-64 shows several ways that various forms of strip materials can be held in place. Two methods predominate: the use of an attachment fin and sidewall friction in a slot. In the sketches on figure 2-65, the gasket shown can be spot-welded directly or held by bonding cement. The bonding compound should be applied in small 1/8-inch diameter droplets, every one or two inches, or as a continuous thin strip on wider materials.

(5) Figure 2-71 illustrates a typical conductive gasketing problem: an equipment panel sliding into a case with an internal flange. If the panel also holds the chassis, then the entire weight of the equipment, exclusive of the case, is borne by the flange. Therefore, it is very desirable that there be a rigid, positive stop on any gasketing material used in the joint between the panel and case flange. The gasket shown is ideally suited for this application. It has an aluminum extrusion to which is attached a resilient gasketing material. The panel comes to a stop at the thickness of the extrusion. Because the uncompressed thickness of the material is larger than the thickness of the extrusion, it operates under pressure.

(6) Often, AN-type connectors are mounted on bulkheads and must maintain both electromagnetic radiation and fluid seals. In such cases, woven aluminum screens, impregnated with neoprene, are suitable. The connectors are made to close tolerances and are usually mounted on a sufficiently rigid surface so that the small amount of compression of this material is adequate. These gaskets are often so small that no other material could be used to make a successful combination sealing gasket.

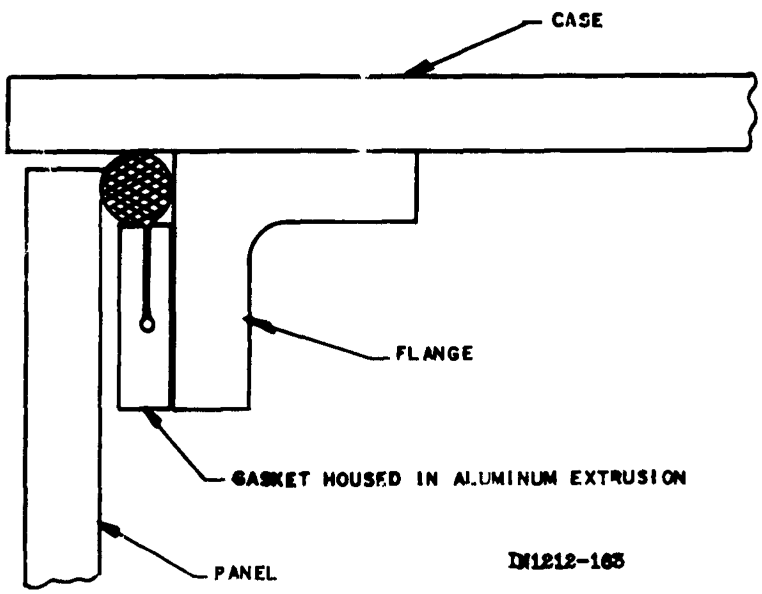



Figure 2-71. Typical Conductive Gasketing Problem

(7) Some applications require seals to be both pressure and interference tight. Figure 2-72 illustrates a typical sample of this type of seal. A conductive gasket is mounted directly to rubber, or other elastic material, to make a combination pressure and interference seal. This combination seal can be held in place by bonding the rubber to one of the surfaces, or by use of the screw holes in the gasket.

(8) In some equipment, there may be a requirement for a pressure differential of only one or two inches of water; and for cooling air to be ducted so as not to leak out indiscriminately at joints. A combination of air-tight and interference-tight gasket material to meet this requirement can be made by knitting one or two layers of mesh over a neoprene or silicone material (fig. 2-73). It will keep rain out, but not water under pressure. It also represents a compromise as an interference shield, as there is much less metal in this structure than in an all-metal gasket.

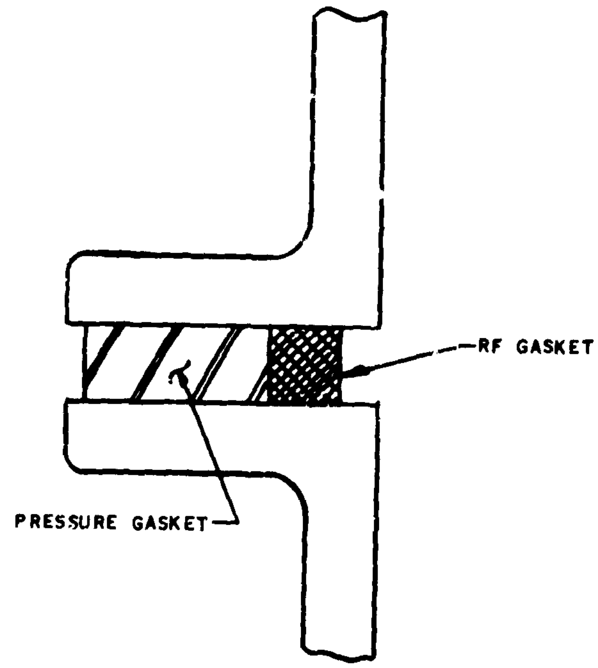



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Figure 2-72. Pressurized Conductive Gasket Application

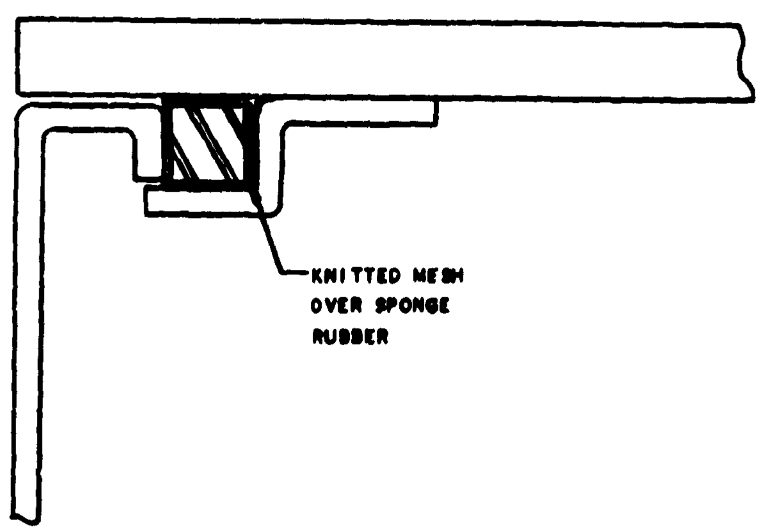


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Figure 2-73. Rain-Tight Conductive Gasket Seal

(9) Gasketing materials, such as woven metal-neoprene combinations, are usually less effective than metal finger or mesh gaskets. The metal-neoprene type requires much greater pressure to achieve satisfactory compression, and requires a greater number of fasteners per unit length. Gasketing materials composed of conductive plastics have recently become commercially available and are especially useful in cases where an air or dust seal is also required. In general, they require greater pressure than metal mesh for satisfactory compression and necessitate strong, well-designed mating surfaces. Such materials should have a resistivity at least as low as 0.001 ohm-centimeters to achieve reasonable shielding effectiveness at frequencies as low as 0.15 mc.

(10) Figure 2-74 depicts insertion loss, in decibels, of a resilient metal gasket, plotted against applied pressure. This insertion loss is the reduction of energy leakage that results when a conductive gasket is inserted in a previously ungasketed joint, all other parameters remaining the same (fig. 2-75). Since $E_2$ is the same in both cases, and $E_1$ can be assumed to be constant, the calibrated attenuator is a direct substitute for the gasket and yields the insertion loss directly. Shielding effectiveness or attenuation should not be used as a figure of merit for conductive gaskets. In some discussions on shielding, shielding effectiveness is defined as the over-all effectiveness of the entire shielding system; attenuation is the shielding capability of the metal of the shield. When tests were run for various combinations of frequencies, gasket materials, and joint materials, the specific values of insertion loss changed, but the shape of the curve remained approximately the same. There was always a point at the knee of the curve where additional pressure did not produce much insertion loss. On the average, this point occurred at 20 pounds per square inch.

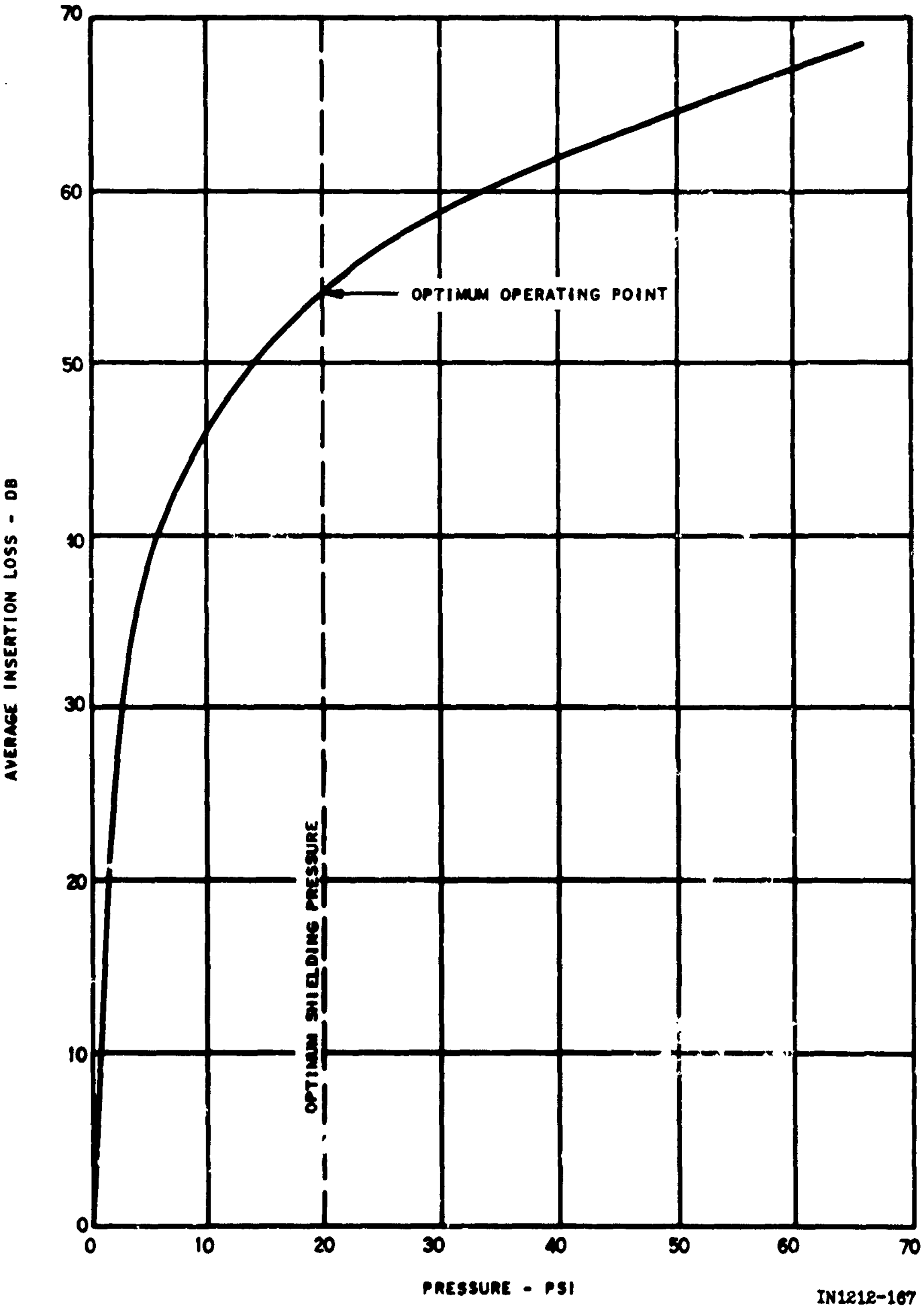
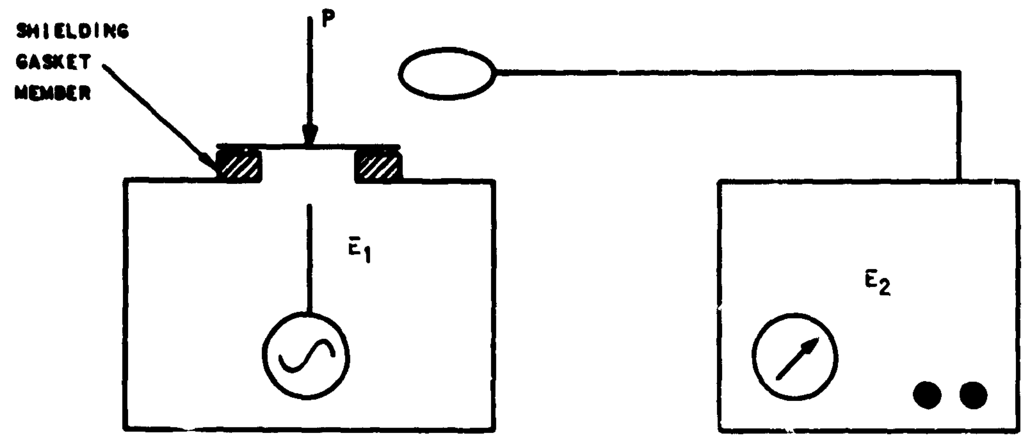
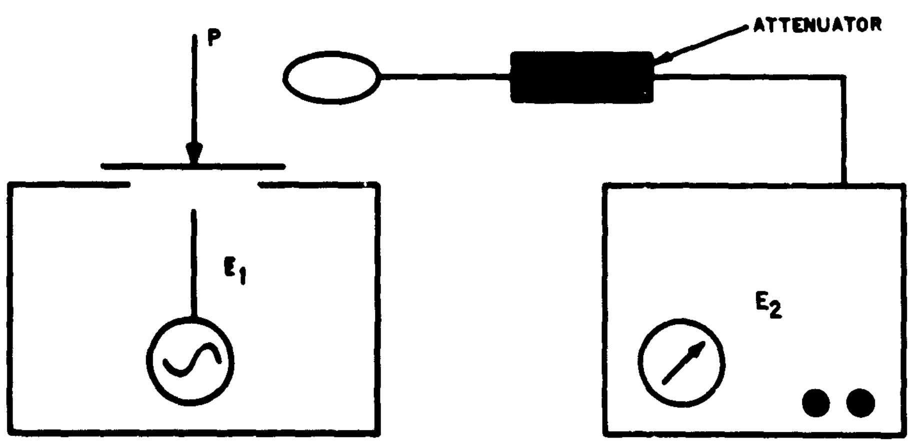



Figure 2-74. Insertion Loss Vs Pressure for Resilient Metal Gasket




A. $E_2$ OBTAINED FOR VARYING VALUES OF P WITH GASKET SHIELDING MATERIAL. $E_1$ HELD CONSTANT.




B. $E_2$ REPEATED FOR VARYING VALUES OF P WITH NO SHIELDING GASKET MATERIAL PRESENT. INSERTION LOSS MEASURED BY ATTENUATOR. $E_1$ HELD CONSTANT.

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Figure 2-75. Technique for Measurement of Insertion Loss of Resilient Metal Gasket

(11) Joints vary not only in their degree of misfit, but also in frequency and manner in which they are opened and closed during the life of the equipment. Most resilient metallic gaskets will take some set when compressed, and will not return completely to their initial state when released. The second compression cycle, then, will not be the same as the first. Therefore, if a gasket is to be used in a joint that will be opened and closed frequently, this effect must be considered. Joints are classified as follows:

(a) Class A, permanently closed. After initial closure, a Class A, permanently closed joint is opened only for major maintenance or repair. Feed-through mounted interference filters and many waveguide joints are examples of Class A joints.

(b) Class B, fixed position. In a Class B joint, the relative positions of mating surfaces and gasket are always the same. For instance, any point on the edge of a door will close against its equivalent point on the jamb after every opening. In a Class B joint, the gasket is compressed and released from the same operating height at the same point on the gasket many times during its life. Hinged lids and doors, and rack and panel installations, are typical of Class B joints.

(c) Class C, completely interchangeable. A Class C joint is one in which mating surfaces and shielding materials are completely interchangeable, and/or the relative positions of two mating surfaces and gasket may change. In a Class C joint, the gaskets may be compressed to several different operating points many times. Symmetrical cover plates and waveguide choke flanges are examples of Class C joints.

c. Gasket Characteristics. Various materials have been used to combine resiliency and conductivity. These characteristics can be combined by using several methods and materials. Some of the more common materials are tabulated in table 2-27 and described in the following paragraphs.

TABLE 2-27. CHARACTERISTICS OF CONDUCTIVE GASKETING MATERIALS

| Material | Chief Advantages | Chief Limitations |
| --- | --- | --- |
| Compressed knitted wire | Most resilient all-metal gasket (low flange pressure required)<br>Most points of contact<br>Available in variety of thicknesses and resiliencies | Not available in sheet form. (Certain intricate shapes difficult to make)<br>Must be 0.040 inch or thicker |
| AEEL gasket or equivalent | Best break-through on corrosion films | Not truly resilient. Not generally reusable |
| Armour Research gasket or equivalent | Combines fluid and conductive seal | Requires 1/4-inch thickness and 1/2-inch width for optimum shielding. |
| Aluminum screen impregnated with neoprene | Combines fluid and conductive seal. Thinnest gasket. Can be cut to intricate shapes | Very low resiliency (high flange pressure required) |
| Soft metals | Cheapest in small sizes | Cold flows, low resiliency |
| Metal over rubber | Takes advantage of the resiliency of rubber | Foil cracks or shifts position. Generally low insertion loss yielding poor rf properties. |
| Conductive rubber | Combines fluid and conductive seal | Practically no insertion loss giving very poor rf properties |
| Contact fingers | Best suited for sliding contact | Easily damaged. Few points of contact |

(1) Knitted wire mesh. Knitted wire mesh material is made of many interlocked loop-shaped springs; it combines springiness with flexibility and cohesion. These properties are retained when the mesh is compressed and make it dense enough to be an efficient shielding material.

(2) AEEL gaskets or equivalent. AEEL gaskets, or their equivalent, are 10-mil strips of beryllium copper and are made by puncturing thin sheets in both directions with a nail. The resultant sharp raised points do an excellent job of making good contact with both sides of the joint. Such a gasket can be imbedded in rubber to give a good pressure seal as well as a good rf seal. It can also be used where surfaces are anodized or corroded because it has the ability to cut through the films, making good metal-to-metal contact.

(3) Armour Research gaskets or equivalent. Armour Research gaskets, or their equivalent, have many wires, each shaped like an open V rotated 90 degrees from the plane of the mating surfaces and imbedded in silicone rubber; they make very efficient conductive gaskets (fig. 2-76). The wires in the gaskets have the ability to puncture any oxide film on mating surfaces; and the bends in the wires allow the gaskets to compress while the ends of the wires establish contact.

(4) Woven aluminum mesh or screen impregnated with neoprene. Woven aluminum mesh or screen impregnated with neoprene is ground off to expose peaks of the aluminum mesh on both sides. One example of this type of material is 16-gauge aluminum screen impregnated with neoprene.

(5) Soft metal. Soft metals, such as copper or lead, have been used as conductive gaskets.

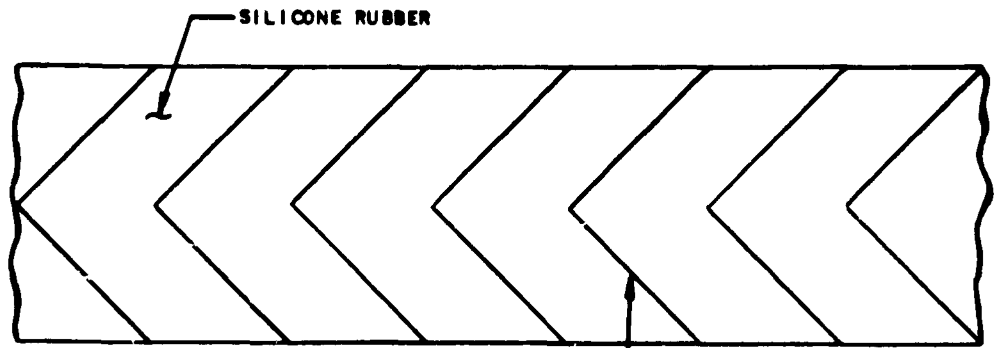



Figure 2-76. Interference Gasket of the Armour Research Type

(6) Metal over rubber. By covering rubber with metal foil, or wire, the resiliency of rubber is combined with the conductivity of metal.

(7) Conductive rubber. Conductive rubber has been tested, but at present does not have sufficient conductivity.

(8) Serrated contacted fingers. Resiliency, in the case of serrated contacted fingers, is achieved by having cantilever springs make wiping contact between the two surfaces. If the fingers are used in this manner, they can be considered adequate conductive gaskets (fig. 2-77).

d. Conductive Gasket Selection. The design goal for conductive gasket application is a gasket, resilient enough to make continuous contact despite irregularities of both surfaces, at a pressure high enough to result in low impedance across the joint. Most conductive gaskets take some amount of compression set. Since the gasket in a Class A joint is compressed only once, compression set is of little concern. In a Class B joint, the set results in reduced uncompressed gasket height. In a Class C joint, maximum compression set may, in a subsequent use of the gasket, occur at the joint of minimum set. This means that gaskets in Class A joints can stand more compression than in Class B or C. Class C gaskets can tolerate the least compression. Table 2-28 is a guide to selection of rf gaskets based on mechanical suitability.

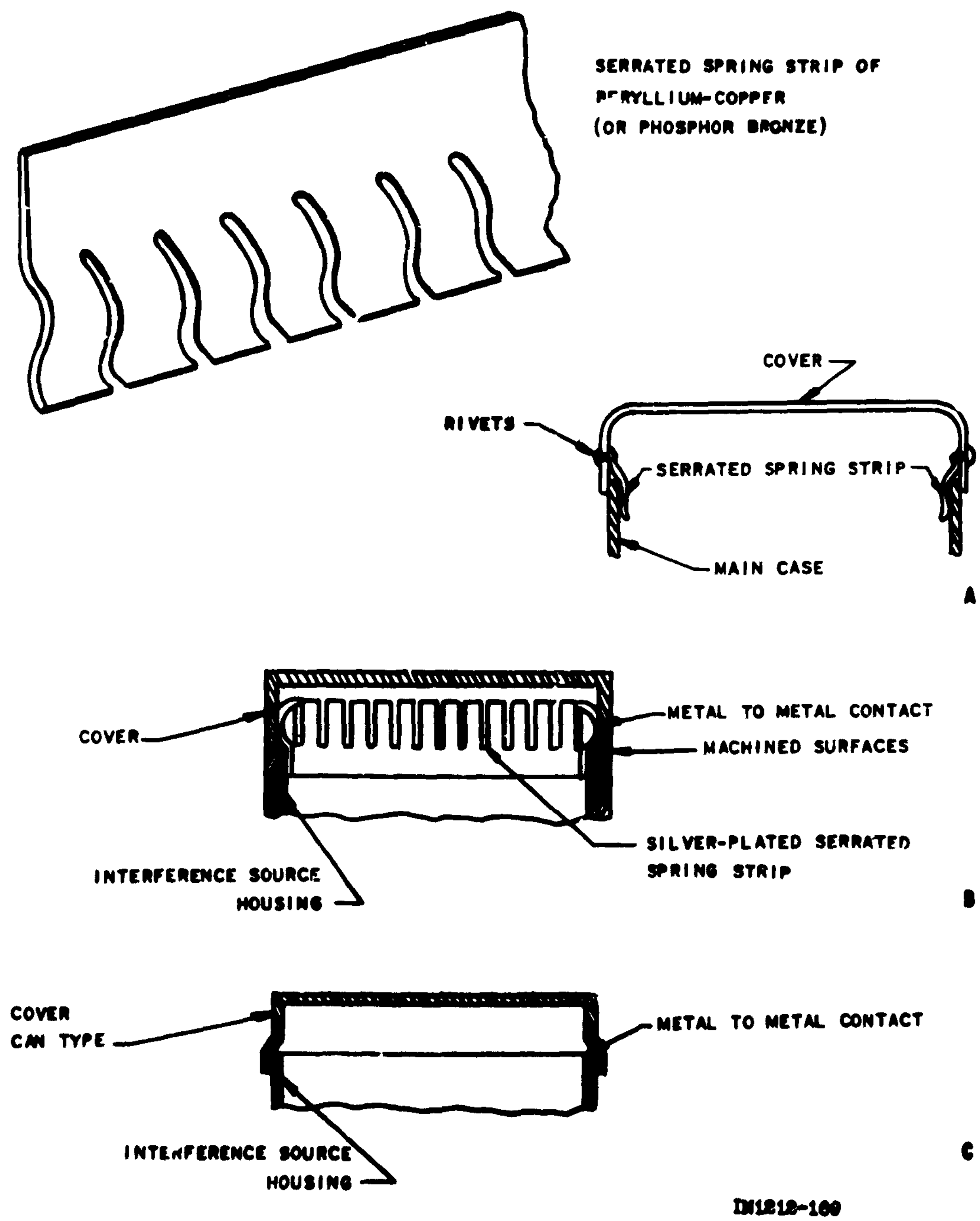



Figure 2-77. Metal-to-Metal Contact Techniques of Cover and Case for Interference Shields

## TABLE 2-28. GUIDE TO CHOICE OF TYPE OF RF GASKET BASED ON MECHANICAL SUITABILITY

1, well suited; 2, can be considered; 3, [illegible]t suited

| | | Mesh Strips: ○ | Mesh Strips: □ | Mesh Strips: ▭ | Mesh Strips: ▭▭ | Extrusion Gasket | Combination Strip | Formed RF Gaskets | AEEL Gaskets or Equivalent | Combination Gaskets | Woven Alum. + Neoprene | Mesh over Rubber | Conductive Rubber[i] | Fingers |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| Attachment Method | Held in a slot or groove by tight sidewall fit | 1 | 1 | 3 | 3 | 3 | 2 | 1 | 3 | 2 | 3 | 1 | 1 | 3 |
| | Nonconductive spot-bonding[a] | 2 | 1 | 2 | 2 | 3 | 2 | 1 | 3 | 2 | 2 | 1 | 1 | 3 |
| | Nonconductive bonding away from rf gasket portion[b] | 3 | 3 | 1 | 1 | 3 | 1 | 3 | 3 | 1 | 3 | 3 | 3 | 3 |
| | Bond rf gasket portion with conductive adhesive[c] | 2 | 1 | 2 | 2 | | 2 | 1 | 3 | 2 | 2 | 1 | 1 | 3 |
| | Screw, spotweld, or rivet | 3 | 3 | 1[d] | 1[d] | 1 | 3 | 3 | 2 | 3 | 3 | 3 | 3 | 1 |
| | Solder | 3 | 2 | 2 | 2 | 3 | 3 | 2 | 2 | 3 | 3 | 3 | 3 | 1 |
| | Position by bolts through bolt holes | 3 | 3 | 2 | 2 | 1 | 1 | 2 | 1 | 1 | 1 | 3 | 1 | 3 |
| | Pressure-sensitive adhesive backing | 3 | 3 | 3 | 3 | 3 | 1 | 3 | 3 | 1 | 3 | 3 | 3 | 3 |
| Other Gasketing Functions | Cooling air tightness | 3 | 3 | 3 | 3 | 1 | 1 | 3 | 3 | 1 | 1 | 1 | 1 | 3 |
| | Rain tightness | 3 | 3 | 3 | 3 | 1[e] | 1 | 3 | 3 | 1 | 1 | 1[e] | 1 | 3 |
| | Pressure tight | 3 | 3 | 3 | 3 | 3 | 1 | 3 | 3 | 1 | 1 | 3 | 1 | 3 |
| Pressure Available | 0 - 5 psi[f] | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 2 | 1 | 3 | 1 | 1 | 1 |
| | 5 - 50 psi | 1/2 | 1 | 1/2 | 1/2 | 1 | 1 | 1 | 1 | 1 | 1 | 1/2 | 1 | 2 |
| | Over 50 psi | 2 | 2 | 2 | 2 | 1 | 2 | 2 | 1 | 2 | 1 | 2 | 1 | 3 |
| Total Joint Unevenness[g] | Less than 0.002 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 |
| | 0.002 to 0.030 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 3 | 1 | 2 | 1 |
| | 0.030 to 0.060 | | | | | 1 | 1 | 1 | 1 | 2/3 | 1 | 3 | 1 | 2/3 | 1 |
| | Over 0.060 | 1 | 1/2 | 1 | 1 | 2 | 2 | 1/2 | 3 | [illegible] | 3 | 1 | 3 | 1 |
| Space Available, Width | Less than 0.060[h] | 1 | 1 | 3 | 3 | 3 | 3 | 2 | 3 | 3 | 1 | 3 | 1 | 3 |
| | 0.060 to 0.500 | 1 | 1 | 1 | 2 | 2/3 | 2 | 1 | 2 | 2 | 1 | 1 | 1 | 2 |
| | 0.500 to 1.50 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 |
| Space Available, Thickness | Less than 0.030 | 3 | 3 | 3 | 3 | 3 | 3 | 3 | 2 | 3 | 1 | 3 | 2 | 2 |
| | 0.030 to 0.060 | 1 | 1 | 2 | 2 | 3 | 2 | 1 | 1 | 2 | 3 | 2 | 1 | 1 |
| | 0.060 to 0.090 | 1 | 1 | 1 | 1 | 2 | 2 | 1 | 1 | 2 | 3 | 2 | 1 | 1 |
| | Over 0.090 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 3 | 1 | 3 | 1 | 1 | 1 |
| Type of Joint | Compression only | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 |
| | Combined compression & sliding | 2 | 2 | 2 | 2 | 2 | 2 | 2 | 3 | 2 | 3 | 2 | 2 | 1 |
| | Sliding only | 2 | 2 | 2 | 2 | 2 | 3 | 2 | 3 | 3 | 3 | 2 | 3 | 1 |

a - Nonconductive spot bonding: A nonconductive adhesive can be used directly under an rf gasket if it is used only in 1/8- to 1/4-inch diameter spots, 1- to 2-inches apart.

b - A nonconductive adhesive can always be used continuously if it is used under the attachment or rubber portion of combination strip and gaskets, but not under the rf gasket itself.

c - A conductive adhesive can be applied continuously under an rf gasket.

d - With backing strip over attachment fins.

e - If mesh-over-rubber version of extrusion gasket is used.

f - Evaluation is only for mechanical suitability. Pressure may be high enough to give sufficient insertion loss.

g - Evaluation based on space being available to use thick enough gasket.

h - Including space for attachment method integral to material considered.

i - Evaluation is not based on electrical suitability of conductive rubber which is generally poorer than other materials listed.

e. Gasket Materials. Various forms of knitted wire conductive gaskets are available in monel, silver-plated brass, and aluminum. These materials are shown in table 2-29. Monel is the most corrosion resistant material of the three; and aluminum is the least corrosion resistant. Frequently, conductive gaskets are used in contact with an aluminum structure; the problem of compatibility then arises. In certain respects, an aluminum gasket would be more compatible with aluminum. However, monel is a better choice because of superior resiliency and corrosion characteristics. The choice of monel is based on its own corrosion resistance. If an aluminum gasket were used, the gasket would, itself, be subject to corrosion. Even though monel may cause some corrosion of the aluminum enclosure, this is a minor consideration compared to gasket loss by corrosion. Monel is also recommended for use with magnesium, aluminum, cadmium-plated steel, stainless steel, and similar metals. While, in the strict sense, monel is not compatible with these metals, its use is permitted because the monel will be present in a much smaller mass, reducing galvanic corrosion. The ranking of conductivity requires some explanation. If only monel, silver-plated brass, and aluminum are considered, the intrinsic conductivity is as shown in table 2-29. The numbers in parentheses in that table indicate that if silver only is considered instead of silver-plated brass, then the ranking would be 3, 1, 2, rather than 3, 2, 1. In conductive gasket work, however, the intrinsic conductivity of the material is not the important consideration. Because surface corrosion films can form and greatly reduce the actual conductivity of a gasket, material should be ranked by its conductivity with surface films. When this is done, they rank 1, 2, 3, as shown, except when silver-plated brass is very clean and fresh; then the ranking is 2, 1, 3. For tensile strength, springiness, and hardness, monel is usually recommended, except that silver-plated brass should be used if maximum conductivity is needed and the silver is to be used in a controlled atmosphere.

TABLE 2-29. GASKET MATERIAL CHARACTERISTICS

| Material | Corrosion | | Conductivity | | Mechanical | | |
|---|---|---|---|---|---|---|---|
| | Intrinsic | In Contact With Aluminum | Intrinsic | Over-all | Tensile | Spring | Hardness |
| Monel | 1 | 2 | 3 | 1 (2) | 1 | 2 | 1 |
| Silver-plated brass | 2 | 3 | 2 (1) | 2 (1) | 2 | 1 | 2 |
| Aluminum | 3 | 1 | 1 (2) | 3 | 3 | 3 | 3 |

1: most desirable

3: least desirable

2-23. Special Shielding Techniques

a. Carbon Arc and Gas Discharge Lamps. Carbon-arc lamps are widely used in projectors and reproduction devices. They produce a continuous electrical arc between two carbon electrodes and are, therefore, very severe interference generators. Because carbon-arc lamps are usually used in small enclosures, and light is beamed in a single direction, partial shielding is feasible and effective. In addition, feed-through or bypass capacitors may be installed in power leads to prevent conduction of interference out of the enclosure.

(1) Gas discharge lamps such as fluorescent lights, ultraviolet, and neon lamps are intrinsic sources of interference; interference control is largely a matter of isolation and containment. This interference arises from the rapid ionization that takes place with each cycle of the line frequency which generates impulsive bursts of wide-band interference. The ionic discharge often generates a single frequency, usually in the low megacycle range, which appears at the terminals of the lamp and is radiated from the glass bulb and interconnecting

leads. Fluorescent lamps contain mercury vapor at low pressure which is ionized by the flow of electrons in the tube. The subsequent deionization causes ultraviolet radiation. This excites the phosphor coating on the inside of the tube, causing it to emit light in the visible region. Since this process essentially employs a continuous arc, it gives rise to radio interference.

(2) There are three ways in which these lamps can transmit interference to receivers: by radiation directly from the lamp, radiation from power leads, or transmittal by conduction through a common power supply system. Suppression of direct lamp radiation is a function of lamp shielding. This approach is not often very effective, however, because shielding can interfere with the normal lighting function. Elimination of most power-line conducted or radiated fluorescent lamp interference is accomplished with the aid of feed-through or bypass capacitors. For systems that employ starters, such capacitors may be placed across starter terminals, as shown on figure 2-78. For starterless systems, the capacitors are mounted in the ballast, as shown on figure 2-79. Most lighting fixtures have these capacitors built in, but this may not be the case when manual starting is employed. For manual starting, a capacitor of from 0.006 to 0.01 $\mu$f should be installed across lamp leads. In cases where it is desirable to reduce interference further in power leads, filters should be used.

(3) Lamps of this type cannot be completely shielded with sheet metal without eliminating all of the useful light output. However, screen wire can be used, but reduction in light transmission by suitable screening may be as great as 50 percent. Progress has been made in the development of coated glass which transmits over 90 percent of the visible light while providing adequate interference shielding. The

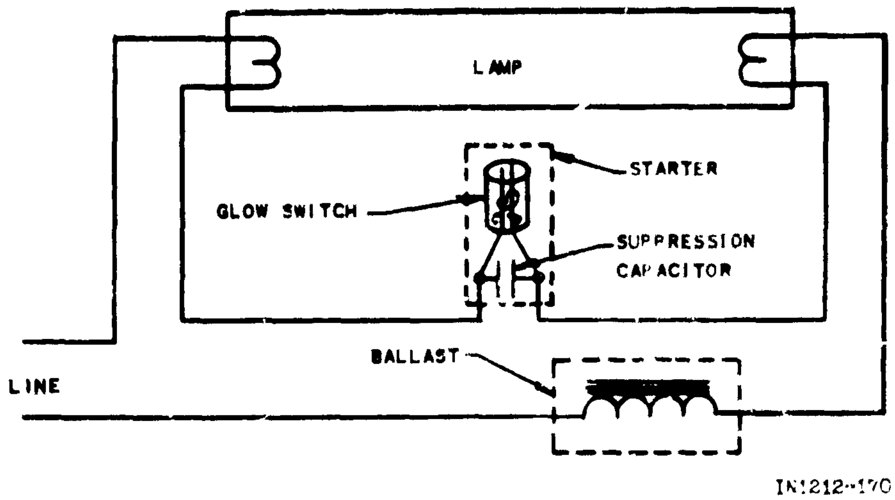



Figure 2-78. Starter Type Fluorescent Lamp Suppressor

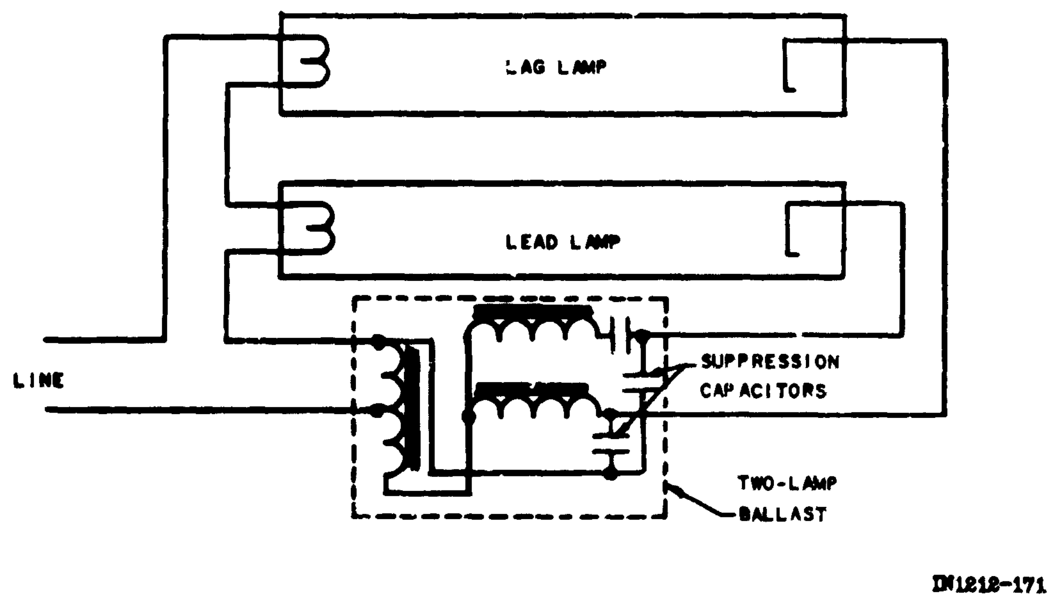



Figure 2-79. Starterless Fluorescent Lamp Suppressors

function of the coating on the glass is to intercept and ground out radiated interference. Typical construction consists of a heat-resistant, borosilicate glass panel having a permanently bonded, transparent, electrically-conducting film applied to its smooth side. A 0.25-inch-wide metal grounding strip is fired onto the film around the periphery of the glass panel. A conductive silver paint is applied to make good contact between the glass panel and the frame; and the frame is bonded to the metal fixture ground plane. The glass coatings usually exhibit resistances ranging from 50 to 200 ohms per square inch and substantially reduce radiated interference in the frequency range from 0.014 to 25 mc.

(4) The standard slimline fluorescent lighting fixtures represent a potential source of conducted and radiated interference. Confirming tests performed on an 8-foot slimline fixture indicated conducted interference of 700 microvolts per kc bandwidth at 0.5 megacycles, and 18 microvolts per kc at 10 mc. Radiated interference exceeded the limits of MIL-I-6181B up through 1 mc. As an example, 700 microvolts per kc broadband interference energy in a circuit having a bandwidth of 20 kc would produce the equivalent of a 14,000 microvolt signal. Further, a large number of such fixtures could generate even higher interference voltage amplitudes, depending on inherent phase relationships. To deal effectively with these levels of broadband interference, waveguide type filter shields of honeycomb material can be successfully employed. Low rf impedance bonding of the shield assembly to the fixture frame and, in turn, to ground through conduits, etc., becomes of utmost importance. The use of capacitor networks (by-pass or feedthroughs) or power line filters (and conduits) are recommended for attenuation of conducted interference phenomena.

b. Magnetic Shielding Techniques. Well-engineered circuits can be seriously affected by interference from spurious magnetic fields set up by neighboring components. Experience has shown that these effects can be minimized by judicious orientation of components on the chassis. There are many instances, however, where this cannot be accomplished because of space limitations. In addition, many low-level applications can be upset by the earth's magnetic field. This problem can be critical in amplifier circuits where the extraneous signal may be amplified thousands of times. For example, the field generated by the driving motor of a tape recorder is of sufficient magnitude to dominate the signal completely unless the record and play back heads are well shielded. The electron beam of a cathode-ray tube can be bent by the field of a transformer used in the circuit. A magnetic compass is of such strength as to affect a radar scope six feet away. Because of such effects, input transformers handling very low-level signals must be protected from any spurious fields. Some relief can be obtained in the design of transformers by using a hum-bucking construction, where some of the fields are bucked out by splitting the windings on the magnetic core structure. Another solution is the use of a core structure in which air gaps between lamination pairs (where fringing of flux develops) are contained within the electrical coil. In general, the solution of most of these problems lies in the use of a magnetic shield around the affected component.

(1) A magnetic shield is actually a low-reluctance path in which the magnetic field is contained. For this reason, shields are generally made of high permeability nickel-iron alloy such as Mu-metal. Cast iron and other relatively low permeability materials have also been used by compensating with a much heavier thickness. For example, most nickel-iron alloy shields are produced in thicknesses of 0.025 to 0.035 inch, while a comparable shield in cast iron would probably require a 0.125 to 0.25-inch thickness. Where weight and size are critical, nickel-iron alloys should be specified.

(2) There are many low-level applications where a single shield will not reduce the field to a low enough value. In such cases, a nest of shields is necessary. Additional shields are placed over the first one to achieve the degree of isolation required. These should be completely separated from each other to ensure maximum permeability of the group and a high reluctance gap between each shield. This is often accomplished by using 0.010-inch thick Kraft paper as a separator.

(3) Structurally, magnetic shields fall into two broad classes: those produced by deep drawing from flat blanks and those formed and welded. Because of difficulties involved in deep drawing nickel-iron alloys, drawn shields are usually confined to smaller sizes; and because nickel-iron alloys work harden very rapidly in the drawing process, generous radii must be provided to prevent tearing. In drawing Mu-metal cans, care should be taken to relieve the internal strains as quickly as possible. Shields for such items as transformers, cathode-ray tubes and photomultiplier tubes are fabricated from flat, unannealed sheets of metal. The material is bent on brakes or rolls, and the joints are overlapped and spot welded. All holes and slots are pierced prior to the forming operation. After fabrication, shields are given the final heat treatment in pure, dry hydrogen at 2050°F to develop the required permeability. Before this final treatment, the permeability is approximately five percent of the ultimate that is developed by the annealing process. For the most critical applications, deviations from this treatment will result in an inferior unit which is apt to have nonuniform magnetic characteristics in addition to low permeability. It has been found that a 3/8-inch overlap of material is sufficient to prevent any penetration by the extraneous field. Spot welding at intervals of 1/2-inch is adequate to seal the joint.

(4) Placement of joints in the shield surface affects the shield effectiveness. In a cathode-ray tube shield, for example, joints in the axial direction of the tube have little degrading effect on shielding effectiveness, but joints normal to the axis reduce the effectiveness of the shielding. Figure 2-80 shows examples of these constructions, and also the developed blanks from which they are formed. The blank used to form the shield with the axial joint (fig. 2-80A), has a maximum flux path; the normal joint in the second example (fig. 2-80B) divides the blank into two parts that have lower permeance, resulting in decreased shielding effectiveness. If space and mounting difficulties are no problem, costs can be cut by eliminating expensive layouts, as shown on figure 2-81. Here, a simple frustum of a cone that follows the contour of the tube is substituted for a three-piece construction (fig. 2-81A). In addition to the economy realized through simpler construction, annealing costs can be greatly reduced by nesting the cones in the annealing box, as shown on figure 2-81B.

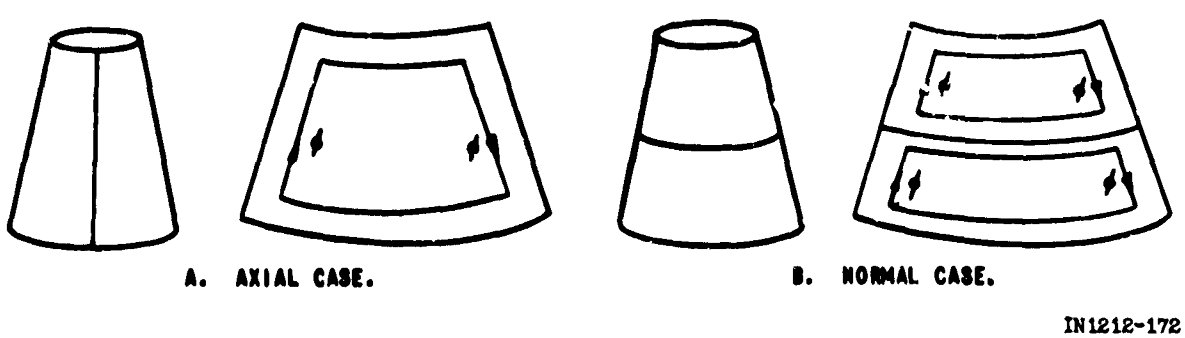
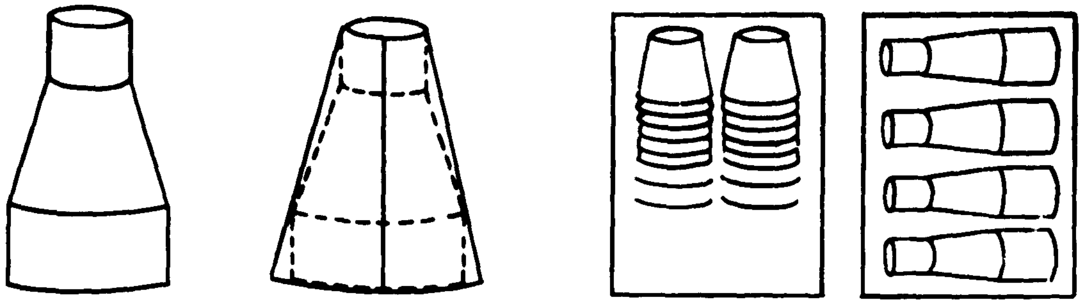



Figure 2-80. Effect of Joint Orientation on Flux Paths



A. CRT SHIELD CONSTRUCTION.

B. NESTING SHIELDS.

IN1212-173

Figure 2-81. Shield Economies

(5) The question is often raised as to the possibility of using annealed Mu-metal sheets to eliminate the much higher cost of annealing in the final form. Unfortunately, the permeability of Mu-metal is adversely affected by cold work of any kind; therefore, the effectiveness of this shield would be inferior to one made in the recommended manner and then given a final heat treatment. This is particularly true because most shields are produced from fairly heavy-gauge material (0.025 to 0.035 inch thick).

(6) For shielding, some of the trade name materials on the market are Mu-metal, Co-Netic AA, Unimag 80, Hi Mu 80, and Hypernom. These materials have generally the same basic constituents of 80 percent nickel, 20 percent iron. The relative shielding effectiveness varies little between materials, but varies a great deal with field frequency. In general, the lower the frequency, the lower the shielding effectiveness. An 80-percent nickel alloy is difficult to draw and form. The best way to handle these alloys is by the polyform process: casting a matrix in a die or mold to the desired shape. The matrix is rotated and cooled while the magnetic shielding material is sprayed on it to the desired thickness. When complete, subsequent grinding, drilling or other machining is carried out;

then the matrix is removed, leaving an enclosure that is adequate for electronic assemblies. The enclosure will have internal dimensions exactly matching the matrix and therefore will be superior to a fabricated can, particularly when sharp corners are required. To achieve optimum efficiency and add to ductility, annealing usually follows this stage. Where necessary, shielding effectiveness may be further improved by laminating. In many cases, a sprayed layer of copper is inserted between layers of shielding material to provide special characteristics. This can be done with ease, at little additional cost. There are cases where it is impossible to add an additional cover because of space and weight limitations. In such instances, polyform is convenient because the shielding material can be applied directly upon existing enclosures of almost any material to the desired thickness and shielding effectiveness.

(7) In general, the solution to the problem of shielding against magnetic fields, internally generated by transformers and coils, takes one of two directions: shielding the source or shielding the component or chassis that may be most affected. Multiple shields are sometimes required to extend the magnetic and electric shielding into higher frequency ranges. Laminates of magnetic and high-conductivity materials are used; the former for shielding of a magnetic field, and the latter for shielding of an electric field. To ensure maximum permeability from these materials, they must be metallurgically prepared with great caution. In addition, alloys of iron-silicon, iron-nickel and iron-cobalt, with proper heat treatment, yield permeabilities higher than that of pure iron.

c. Data Output Openings. Data output openings, such as those for direct view storage tubes, cathode-ray tubes, and meters, represent a large discontinuity in an equipment case. Meters and other visual readout devices often present difficult interference protection problems, particularly at high frequencies, because, at present, materials that are both optically transparent and conductive are ineffectual shields. If fine knitted mesh or conductive glass is used for display shielding, it may not furnish the degree of attenuation required. In most cases, this problem is overcome by installing a shield around the rear of the readout device and filtering or by-passing all leads entering and leaving it, as shown on figure 2-82. A number of semitransparent shielding materials are available for application to data output and display devices. Included are copper mesh screening, perforated metal, conductively coated glass, and conductively coated plastic. The optical and electrical transmission properties of some of these materials are summarized in table 2-30. A comparison of relative levels at three frequencies indicates that, for virtually all of the conductively coated materials, the ratio of electrical transmittance to optical transmittance is relatively high. For instance, a 30-micron-thick gold film on plastic yields an electrical transmittance of 16 percent (at 5.9 kmc) and an optical transmittance of 24 percent. The electrical transmittance decreases with frequency; at higher frequencies, considerably higher electrical transmittance (poorer shielding qualities) can be postulated. The twenty-mesh copper screen, while showing the same order of electrical transmittance as the gold film at 5.9 kmc, has approximately twice the optical transmittance. In addition, the shielding effectiveness of copper mesh improves with decreasing frequency, although at higher frequencies, comparison of the two materials would favor the copper mesh. Economically, copper mesh is more practical because deposition of thin metal films is an expensive process compared with utilization of copper mesh.

(1) Direct-View Storage Tubes. The application of magnetic shielding to direct-view storage tubes practically eliminates the effects of stray magnetic fields. A double shield is required to stop these fields effectively. It may consist of one shield of Netic-type alloy and one shield of Co-Netic-type alloy. Principal Netic-type alloy material characteristics are high-flux

capacities and extremely low retentivity. Co-Netic-type alloy materials are more effective for low-intensity, low-frequency problems. The inner shield should be of Co-Netic type material or an equivalent; the outer shield, Netic type material or an equivalent. The Co-Netic-type shield should be 0.025 inch thick; the Netic-type shield, 0.062 inch thick. In this arrangement, the dc magnetic field is reduced by a factor of 1000, and the ac magnetic field, tested at 60 cycles, is reduced by a factor of 30,000.

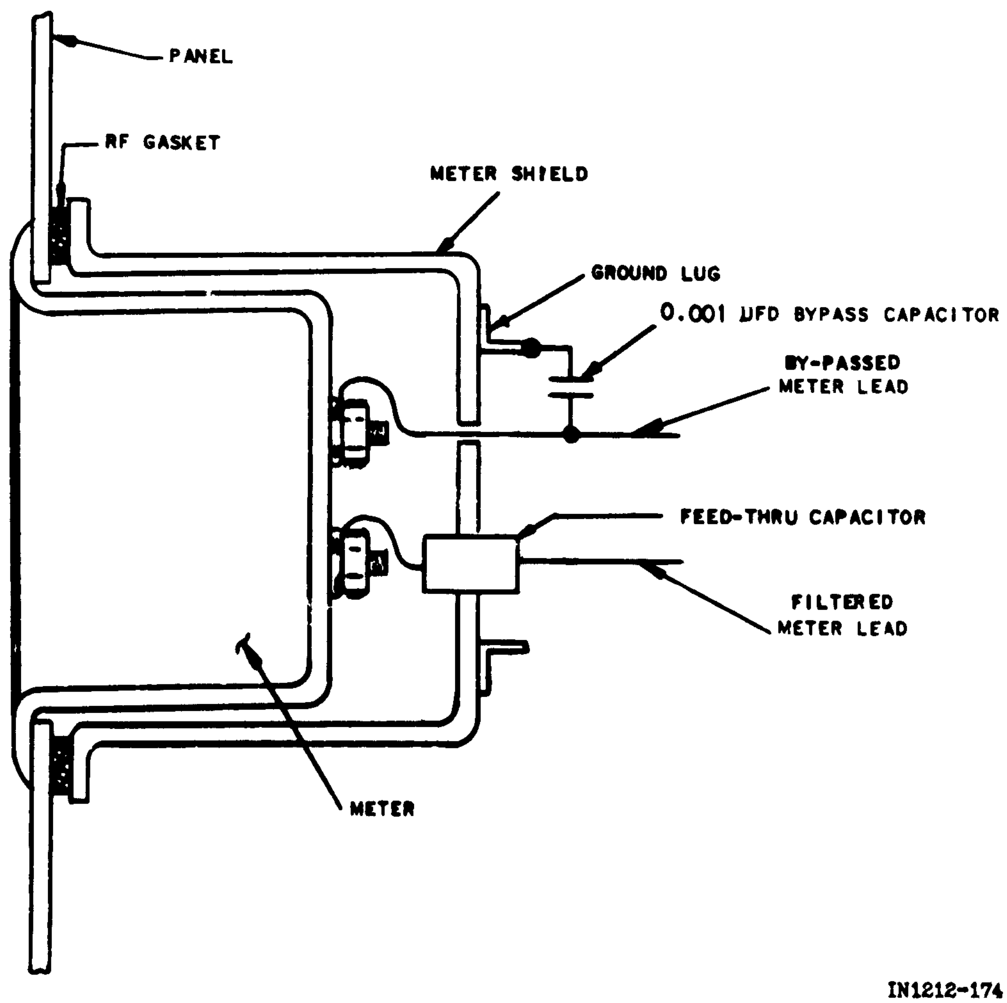



Figure 2-82. Meter Shielding and Isolation

TABLE 2-30. MICROWAVE AND OPTICAL PROPERTIES OF SEMITRANSPARENT SHIELDING MATERIALS

| Material | Microwave Transmittance (Percent) | | | Optical Transmittance (Percent) |
|---|---|---|---|---|
| | 5.9 kmc | 9.7 kmc | 18.8 kmc | |
| Gold film about 11 μ thick on plastic (300 ohms/square) | 23 | 10 | 0.8 | 49 |
| Gold film about 30 μ thick on plastic (12 ohms/square) | 0.16 | 0.1 | 0.01 | 24 |
| Gold film about 75 μ thick on glass (1.5 ohms/square) | 0.04 | 0.01 | 0.004 | 3.2 |
| Copper mesh (20 per inch) | 0.1 | 0.2 | 0.2 | 50 |
| Copper mesh (8 per inch) | 1.0 | 1.3 | 2.5 | 60 |
| Lead glass (X-ray protective, 1/4 inch thick) | 30 | 25 | 16 | 85 |
| Lucite (3/16 inch thick) | 80 | 50 | 25 | 92 |
| Libby-Owens-Ford Electrapane glass, with conductive coating about 150 μ thick (120 ohms/square) | 16 | 16 | 16 | 85 |
| Libby-Owens-Ford Electrapane glass, with conductive coating about 300 μ thick (70 ohms/square) | 9 | 10 | 8 | 80 |
| Corning heating panel glass, with conductive coating about 1.5 μ thick (15 ohms/square) | 1.6 | 1.2 | 0.08 | 45 |

(2) Cathode-Ray Tubes. Cathode-ray tube openings represent one of the largest discontinuities and are of a type most difficult to treat. The disruption of the shield is such that, if not considered in the initial design stages, it may become extremely difficult to achieve an acceptable final product. Treatment of the opening is made additionally difficult by the requirement for unrestricted transparency (fig. 2-83). For good shielding construction, it is necessary for all items that penetrate the shielding, such as pipes and conduits, to be electrically bonded to the shielding at the point of entrance by soldering, brazing, or welding. Handles, latches, screw heads, nails, and other metal projections that pierce the shield should be brazed or soldered to the shield; all breaks should be bonded in continuous seams. These precautions prevent the antenna effect: a metal element that projects through the shielding can act as a receiving antenna on one side, picking up radiated energy and reradiating signals in the opposite direction on the other side. At high frequencies, such isolated hardware is comparable to, and can radiate as, a waveguide probe. Minimal impedance between such a projection and the shielding will eliminate the antenna effect.

(3) Indicating and Elapsed Time Meters. Meter movements, when shielded, are not affected by ac or dc magnetic fields. The most common sources of interference to unshielded meters are transformers, motors, generators, current-carrying cables or buses, solenoids, and other components producing magnetic fields. Shielding against interference from such items permits unrestricted use of meters in otherwise difficult environments.

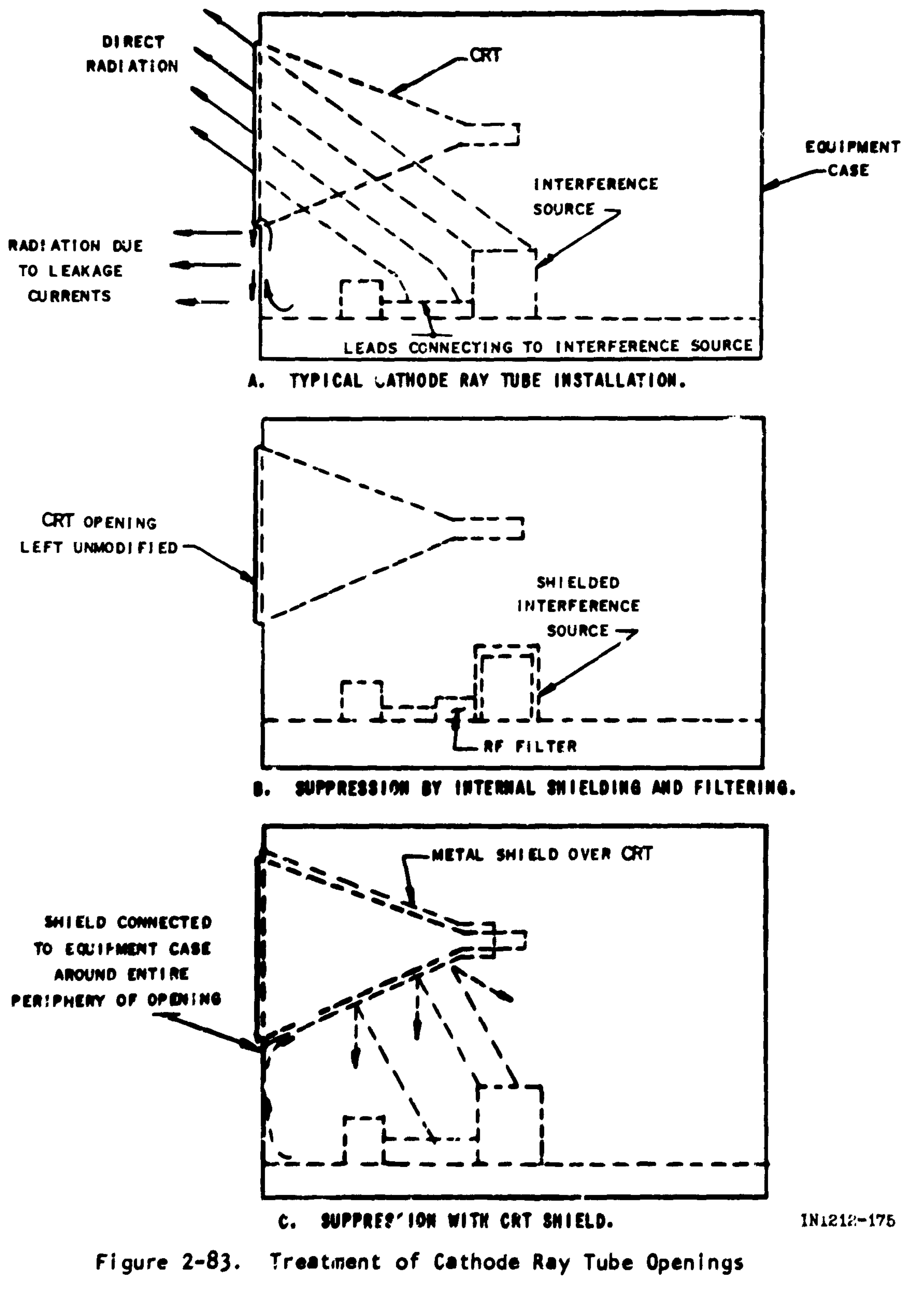



Figure 2-83. Treatment of Cathode Ray Tube Openings

(4) Fuse Holder and Indicator Lamp Openings. In electronic equipment, both active and spare fuses may be sources of radiation because they can act as antennas. The lack of shielding in most fuseholders permits internal high-frequency interference to propagate through the opening in an otherwise well-shielded panel. One solution is to group all fuseholders together; a shield of solid metal with wire mesh gasketing may then be used to surround the fuse cluster. This approach can be utilized for indicating lamps, provided screening or special conducting glass is substituted for the solid-metal fuse holder shield.

(5) Switching Devices. Components, such as solenoids or other devices involving high inrush currents or incorporating switching devices that normally develop high amplitude transients, can prove a source of difficulty in an interference-free design, particularly where space is at a premium. If units are completely enclosed in a shield, the most likely cause of interference radiation will be interconnecting wiring. Mu-metal cannot be drawn into tubing because of the loss of shielding effectiveness resulting from the cold working involved; therefore an adequate shield is often developed by wrapping a continuous layer of annealed tape around the cable. A typical application may involve shielding a cable of approximately 0.5-inch diameter, which has to be flexible in the final assembly. Annealed Mu-metal tape 0.001-inch thick and 0.25-inch wide wrapped in two layers would prove a suitable solution to this problem. The first layer can be spaced approximately 0.125-inch between convolutions, with the second layer overlapping the first layer to cover the gaps between turns. The assembly can be covered with a protective rubber coating and may be flexed without losing its shielding effect. A form of shielded cable using four counterspiral-wound bands of foil, Netic, Co-Netic or their equivalent, is also recommended. This construction is shown on figure 2-84A. The strips can be f[illegible]

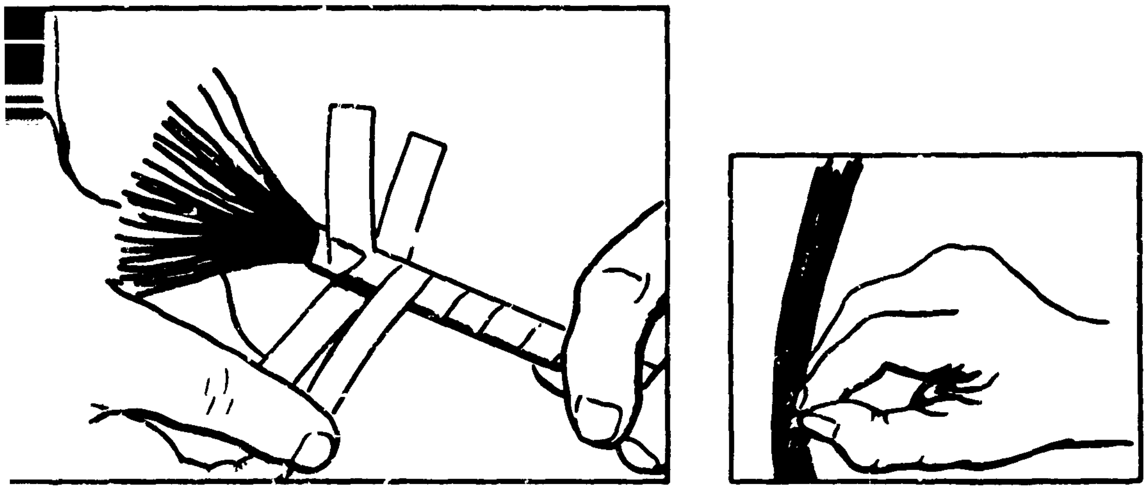



Figure 2-84. Magnetic and Electrostatic Cable Shielding

0.25-inch to 1-inch wide. To minimize leakage between unavoidable gaps, it is necessary to wind the material so as to permit spiral positioning along the length of the cable, with each following layer consisting of another spiral in the opposite direction. Successive layers of the tape would, in this manner, ensure a minimum of gaps and permit flexibility. Such spiral-wound shielded cables are commercially available. A design engineer, who encounters the need for a shield of this nature, can procure the tape in foil form and, for evaluation purposes, fabricate a prototype shield for his own cables. A total of four wraps, or multiples of four, may be necessary for cables carrying appreciable current. For conductors carrying currents greater than two amperes, the first two layers should be Netic S3-6 foil or its equivalent; the remaining layers should be Co-Netic AA foil or its equivalent. Zipper tubing can be used as an efficient means of mechanically holding the foil wraps in place, as shown on figure 2-84B. Zipper tubing is not recommended for cable shielding. Netic and Co-Netic type foils, or their equivalents, are available from 0.002 to 0.007 inch in thickness and in various widths. Also, these alloys provide a simple method of shielding transformers and small reactors; the foils are carefully wrapped around them with the necessary number of layers to provide the desired attenuation level. After wrapping, the cable can be potted or encapsulated to prevent unraveling of the foil. The Netic-type foil and its equivalents exhibit the ability to carry relatively high flux densities; the Co-Netic-type foil and its equivalents have the property of offering maximum attenuation. When an interfering field is of sufficient intensity to partially saturate the Co-Netic-type foil, and therefore limit the realizable attenuation, Netic-type foil can be placed between layers of Co-Netic foil. The Netic-type foil, under these circumstances, acts as a buffer for the Co-Netic-type foil. These foils are extremely thin and cannot function effectively in fields that approach their saturation levels. The extreme versatility of these materials should help the engineer to develop many applications where their use will save space, improve signal-to-noise ratios, and considerably reduce interference.

(6) Conductive Surface Coatings. Eccoshield and its equivalents are highly conductive surface coatings specifically formulated for shielding applications. They are fine, silver-based lacquers that adhere excellently to metal, plastic, ceramic, wood, and concrete. When applied to a nonconductor, the surface resistivity is substantially less than 1 ohm/square inch. Successive coats further decrease the surface resistivity. Eccoshield type ES, or its equivalent, can be brushed on or sprayed. These coatings are often used to improve the rf integrity of metal housings or screen rooms. This is done by applying a surface coating to joints, seams, and contacting surfaces. The material is sufficiently fluid so that it

readily flows into cracks. Metal-to-metal contact is improved significantly by applying such coatings. The surface coating fills slight irregularities and makes such intimate contact with exposed metal that even corroded joints can be made into greatly improved interference seals. Spraying of the coating can drive the silver particles into crevices and other hard-to-reach places. For dilution, or as a cleanup solvent, methyl ethyl ketone (MEK) may be used. In a typical shielded box or enclosure of complete metal construction, application of Eccoshield ES, or its equivalent, to all contact surfaces improves insertion loss by a value of about 30 db over the frequency range from 15 kc to 10,000 mc. Improvement is greater at the high-frequency end of this frequency spread. Thin coating material such as Eccoshield ES will not fill gross voids at joints or seams: a caulking compound such as Eccoshield VX, or its equivalent, must first be used in such applications. When these coatings are to be used on a metal surface, grease, oil, wax, paint, dirt and other nonconductive films must first be removed with a solvent or cleansing agent, or by grinding, buffing or machining the surface to be coated. When electrical contact is established with the metallic base, the coating is applied to the prepared surface by brush or spray The coating should air dry to less than one ohm per-square-inch surface resistivity in one hour. Additional coats can be applied as desired Some of the conductive surface coating and caulking applications are illustrated on figure 2-85 and summarized in table 2-31. The conductive caulking compounds have the consistency of putty and can be applied by hand or with an air-activated gun. The thickness of deposition is easily controlled. Some of the compositions do not harden or set and therefore do not permit joint break but do allow flexibility; others cure to a hard resin surface and, still others, to a rubbery consistency. Elevated temperature cures can be used and are recommended. The joint should be maintained under pressure. Excess compound, squeezed from the joint, can be recovered and reused. Conductive caulking compounds have some adhesivity but should not be relied upon to hold a joint together. Most of these compounds are not cements; however, there are available conductive epoxys with great holding power.

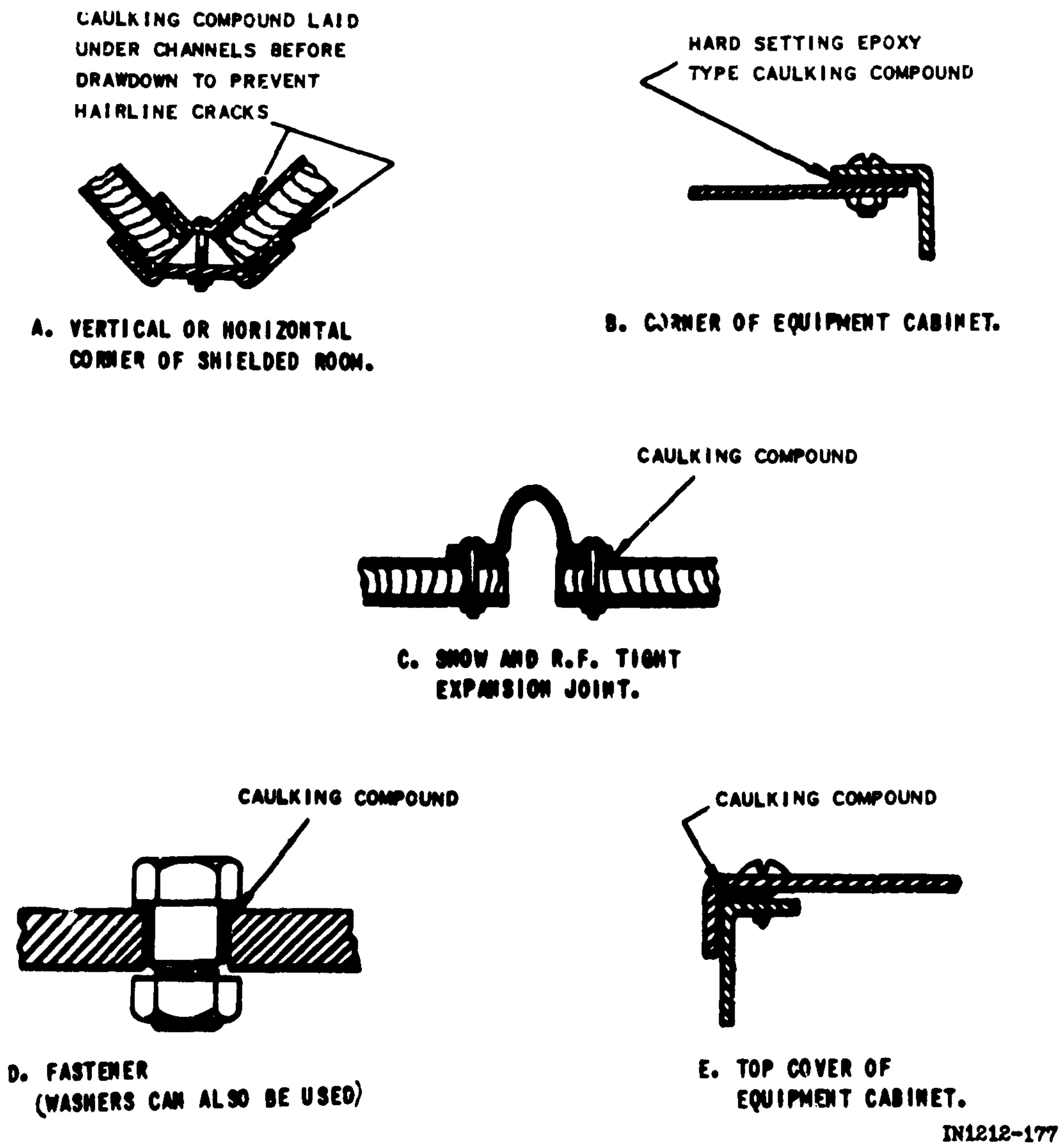



Figure 2-85. Typical Conductive Caulking Applications

TABLE 2-31. CONDUCTIVE SURFACE APPLICATIONS

| Application | Previous Method | Present Method | Advantages |
|---|---|---|---|
| Seams and static joints in rooms, containers, and enclosures | Welding or soldering | Caulking compounds | Ease of application; reliable performance; no damage to metal surfaces, flex resistance; no hairline cracks |
| | Knurled surfaces on clamped joints; metal strips for overlapping seams | Caulking compounds | Bolt spacing increased; no hairline cracks; lighter weight joints; accommodates expansion, contraction, warping, and vibration; weathertight seam |
| Sealing of doors, access hatches, container covers, and flanges | Woven wire mesh strip; copper finger stock | Gaskets<br>Strip gaskets | RF and pressure seal with a single material; low-sealing pressures; small-gap spacings; good compression set; high attenuation |
| | Oriented wire mesh embedded in rubber strip | Strip gaskets | Reuseable; reliable performance; high attenuation |
| | Machined metal seal with molded O-ring, knurled contact surface; wire screen filled with rubber | Seal gaskets | No scarring of flange faces; lower cost; reuseable; low-sealing pressures |
| Sealing of fasteners, bolts, nails, screw heads | Welding or soldering | Seal caulking compounds<br>Seal gaskets | Ease of application; reliable performance; no damage to metal surfaces; easily disassembled. |

## Section V. CABLING

### 2-24. General

Interference may be transferred from one circuit or location to another by interconnecting cabling. The interference may be radiated from a cable or transferred into a cable from external fields. Once interference has been transferred by radiation or common-impedance circuit elements into a cable circuit of an electronic or electrical complex, it can be conducted through interconnecting cables to other elements of the complex. Also, because of cable proximity in cable runs or elsewhere, intra- and/or inter-cable crosstalk may occur as a result of electromagnetic transferrence between cables. The term cabling here encompasses proper selection, assembly, and routing of cables, connectors, and interconnecting circuitry. The cables may be of the commercially available type, prefabricated or specially assembled from a group of individually insulated conductors.

### 2-25. Cable Shielding

Electrical cables may be unshielded, individually shielded, shielded as pairs or groups, or shielded as a whole by a single shield. When a shield is used, it generally consists of a braid or conduit. The purpose of such shielding is to attenuate radiated interference. Proper care should be taken during installation to ensure that shielding integrity is maintained. The choice of cable to be utilized depends upon the characteristic impedance desired, amount of signal attenuation permitted, environment within which the cable must exist, and characteristics of the signal to be transmitted. Signal isolation between different cable circuits is a function not only of cable shielding and the character of signals being distributed, but also of the physical separation of high-level and low-level circuits. The effectiveness of a shield is a function of the conductivity of the metal, contact resistance between strands in the braid, angle and type of weave, strand sizes, percentage of coverage, and size of openings. Multiple layers of shielding, separated by dielectric material (except at connectors), are much more effective than a single layer of shielding.

Although leakage power may be a very small percentage of transmitted power, if the power being carried by a cable is large, this percentage might represent a considerable interference problem to low-level circuits. If a nearby circuit is resonant at or near the transmitted frequency, a small value of transmitted power may give difficulty.

2-26. Cable Types

a. The principal types of cables that are available include unshielded single wire, shielded single wire, shielded multiconductor, coaxial, unshielded twisted-pair, and shielded twisted-pair. Cables are available with both single and multiple shields, in many different forms, and with a variety of physical and electrical characteristics. Proper selection and application of appropriate cables for specific design requirements are highly important in preventing, controlling, and eliminating interference.

b. Shielded cables include the following varieties:

1) Single conductor, single shield

2) Double conductor, single shield

3) Double conductor, each conductor individually shielded

4) Double conductor, each conductor individually shielded and insulated with an outer single or double shield

5) Twisted pair, single shield

6) Twisted pair, double shield

7) Twisted 3-conductor, with single or double outer shield

8) Multiconductor shielded

9) Coaxial single shielded, double shielded, noise-free coaxial, aljak (aluminum jacketed), and heliax

10) Triaxial

11) Cables routed through solid metallic conduit

12) Armored over the shield and jacket

c. Cables are generally specified or identified according to:

1) Size

2) Characteristic impedance

3) Attenuation

4) Shielding (single, double, triple)

5) Power rating

6) Maximum operating voltage

7) Type of jacket (standard black, standard gray, low-temperature black, polyethylene, fiberglass, or armor)

8) Type of dielectric (for example, polyethylene or teflon, solid or spiral ribbon, pressurized or unpressurized)

2-27. Frequently Used Cables

a. Low-Temperature Black Jacket Cable. The low-temperature black jacket cable was developed for low-temperature applications. The cable is black in color, with white ink markings or conventional impression markings, and is suitable for operation between temperature limits of -40°C and +80°C. It is relatively weatherproof and impervious to most mechanical or electrical injury or abrasion.

b. Gray Vinyl Jacket. Gray vinyl jacket cable is similar to black jacket cable except that it utilizes a resinous plasticizer jacket. This cable is suitable for operation between temperature limits of -25°C and +80°C.

c. Polyethylene Jacket Cable. Polyethylene jacket cables are usually thin in jacket cross section; they are used mainly for indoor applications. They are not recommended for outdoor use because of poor abrasion resistance, and because polyethylene is unstable under ultra-violet light. Often, the polyethylene jacket is pigmented dark brown to filter out the sun's rays. Some commercial cables utilize heavy jackets of brown polyethylene which give added abrasion resistance as well as resistance to the sun's ultra-violet rays. This heavy-type cable is suitable for operation between temperature limits of -40°C and +80°C.

d. Fiberglass Jacket. A fiberglass jacket is braided around the cable. It was developed because low-temperature vinyl and polyethylene plastic jackets cannot be used for high-temperature teflon dielectric cables. The fiberglass braid is impregnated with four coats of high-temperature silicone varnish to seal it against moisture and to keep the braid from fraying. It is suitable for operation between temperature limits of -55°C and +250°C.

e. Special Noise-Free Cable. Special noise-free cables fulfill the need for coaxial cables that remain electrically neutral under conditions of shock and vibration. The reduction of inherent noise is achieved for these cables by application of a semiconductive coating over the dielectric.

f. Aljak. Aljak cable has a seamless, extruded, noncontaminating aluminum jacket swaged over either teflon or polyethylene dielectric instead of the usual vinyl jacket. Aljak features lower attenuation, smaller outside diameters, and more than 30% less weight than corresponding RG-/U types. Aljak is frequently used in applications requiring a permanent run of lightweight, rugged, completely weatherproof coaxial cable.

g. Heliax. Heliax is a gas-pressurized, large-diameter, flexible, air dielectric coaxial cable having high power-handling capabilities coupled with very low loss and low VSWR. It has a spiral polyethylene dielectric strip wrapped around a hollow inner conductor tube, and an outer conductor consisting of a flexible metal sheath. The outer sheath conductor is formed of corrugated copper-clad steel, which is first wrapped around the inner conductor assembly and then welded. The outer conductor is covered with tar and crepe filler and enclosed in a vinyl jacket. The high efficiency of heliax cable results from special spiral polyethylene insulation construction; this construction permits a high percentage of air in the space between inner and outer cable conductors and results in an average dielectric constant of less than 1.2. Because the cable is produced by a continuous process, there are no joints. The inner conductor, outer conductor, and spiral insulation are continuous througtout the cable length -- resulting in low VSWR and low insertion loss. The insertion loss of a typical 100-foot length of heliax ranges from 0.7 db at 30 mc, to 5 db at 10,000 megacycles.

h. Triaxial Cable. Triaxial cable is manufactured by placing a high-quality shield-braid over the jacket of standard coaxial cable, and then extruding a full thickness of vinyl jacket over all to provide necessary weatherproofing. An important use for triaxial cable is in test signal distribution systems, where it is necessary to provide additional shielding to minimize cross-cable interference.

i. Metal Armor Cable. Metal armor is used over cable jackets for certain applications where the abrasion resistance of vinyl is insufficient, or where additional cable protection is required because of environmental factors.

j. Commercial Coaxial Cables. Commercial coaxial cables of the RG/U series, most frequently used in the 50-ohm characteristic impedance range, are; RG-8A/U, single shield; RG-9B/U, double shield; RG-177/U, double shield; RG-59B/U, single shield; and RG-55B/U, double shield. For short, low-power, interconnecting equipment cables where line loss is not important, small RG-55B/U and RG-59B/U cables should be used. Where high radio frequencies or medium powers are involved, and/or where line losses become important (such as within extremely low-level circuits, within antenna circuits, or within long cable runs), medium-sized RG-8A/U and RG-9B/U cables should be used. Where high powers are involved, or extremely low-line losses are important, the large-sized RG-177/U cable should be used. For some interconnecting equipment applications, particularly for pulse transmission, it is necessary to use a coaxial cable with low capacitance and high impedance. In such instances, the small-sized

RG-62B/U cables, RG-71B/U cables, or medium-sized RG-63B cables, which have a characteristic impedance in the 93-ohm range, should be used.

k. Flexible Conduit. Flexible conduits for high- and low-voltage shielding usually consist of flexible metal hoses over which are wound one or more layers of braid. Nonconducting coverings are sometimes used over the braid. These coverings provide watertightness and/or added mechanical protection. If applied tightly, they may decrease contact resistance between wires comprising the braid, thereby improving shielding effectiveness. Such coverings should be reasonably rugged and not subject to physical and chemical attack by substances with which they come into contact. They should maintain their desirable characteristics over the anticipated range of operating temperatures. Shielded conduit is used for many diversified purposes, such as:

1) To shield wires and cables electrically that would otherwise radiate interference

2) To provide a channel through which wires and cables may be pulled or pushed for installation or replacement in inaccessible places

3) To protect insulated wires and cables against mechanical damage, for example, chafing and abrasion

4) To keep foreign matter (moisture, oil, grease, gasoline) away from electrical conductors or their insulation

5) To facilitate dissipation of heat for protection of insulation

To be effective, a flexible shielding conduit should be:

1) An effective shield against electrical interference over the entire range of frequencies under consideration

2) Reasonably flexible and capable of being bent to a small radius

3) Rugged enough to withstand considerable abuse and prolonged vibration without serious impairment of either its electrical or mechanical properties

4) Watertight and airtight. The coverings used with it should be immune to attack from lubricants, coolants, antifreeze and fuels

5) Capable of withstanding ambient temperatures likely to be encountered

2-28. Cable Connectors

a. Cable connectors are made in many styles for a multitude of power, signal, control, instrumentation, transducer, audio, video, pulse, and radio-frequency applications. They are made to fulfill special functions, and may be required to be hermetically sealed, submersion proof, and weatherproof. They are manufactured in the straight type, angle type, screw-on type, bayonet twist and lock type, bayonet screw-on type, barrier type, straight plug-in type, and push-on type (table 2-32).

2-29. Cable Application

a. The choice of cable is dictated by the operating signal or power level, frequency range, susceptibility level, and physical isolation. While it is not feasible to set specific rules for cable selection without making an analysis of signal levels and waveforms, the following general rules are suggested:

1) Use unshielded wire for external power circuits (such as 115 vac, 28 vdc)

2) Use shielded wire for multiple-ground, audio frequency, or power circuits

3) Use twisted-pair for audio-frequency circuits grounded at a single point and for internal power circuits

4) Use shielded twisted-pair for single-point ground circuits and multiple-ground circuits where maximum low-frequency isolation is required

TABLE 2-32. CONNECTOR APPLICATION SUMMARY

| Connector Series | Coaxial Cable Size | For RG-/U Cables | Disconnect Style | Voltage Rating | Characteristic Impedance | Freq. Range | Method of Assembly |
|---|---|---|---|---|---|---|---|
| N | Medium & Large | 5,6,8,9, 10,11,12, 13,14,17, 18 | Screw-on type | 500 V peak | 50 ohm 70 ohm (constant) | up to 10 kmc | Manual |
| GR-874 | Medium & Large | 8,9,29,55, 58,58A,59, 62,116 | Push-on type | 1500 V peak | 50 ohm | up to 7 kmc | Manual |
| C | Medium & Small | 8,9,10,12, 14,55,58 | Bayonet Lock type | 1000 V peak | 50 ohm | --- | Manual |
| UHF | Medium & Small | 8,9,10,11, 12,13,55, 58,62,63, 65,71 | Screw-on type | 500 V peak | (nonconstant) | up to 200 mc | Manual |
| LC | Large | 17,18 | Screw-on type | 5000 V peak (modified to 10 kv) | 50 ohm | --- | Manual |
| BN | Medium & Large | 8,9,10,17, 18 | Screw-on type | 5000 V peak | 50 ohm (constant) | --- | Manual |
| BN | Small | 55,58,59, 62,71 | Screw-on type | 250 V peak | (nonconstant) | up to 200 mc | Manual |
| BNC | Small | 55,58,59, 62,71 | Bayonet lock type | 250 V peak | 50 ohm (constant) | up to 10 kmc | Manual & Crimp-on |
| Subminiature | Subminiature | 174 | Screw-on & Push-on types | --- | 50 ohm 75 ohm (constant) | --- | Crimp-on |

5) Use coaxial cable for transmission of rf pulses, high-frequency applications, and where impedance match is critical

6) All twisted circuits should be single-point grounded

Single-conductor, single-shield cable should be used for low-frequency instrumentation applications utilizing ground circuits. It is effective when signals to be transmitted are of moderate levels, and a good low-impedance system ground is available. The double-conductor, single-shield cable is frequently used for single-ended, low-frequency instrumentation applications. It is effective at low signal levels and does not require as good a system ground as does the single conductor cable (as long as it is utilized single ended).

b. Use of one of the two wires of a twisted-pair cable for a circuit return lead affords a reduction in interference radiation because the low-frequency fields produced by the two leads cancel each other. The degree of cancellation is greater for twisted-pair cable than it is for straight double conductor. Twisted-pair conductors, both shielded and unshielded, are effective in reducing interference from magnetic coupling. For example, twisting ac power distribution leads reduces the magnetic field surrounding the wires, thus reducing pickup in circuits lying within the field. Where a signal loop is linked by an interfering magnetic field, pickup is reduced by twisting the signal wire with the ground return wire. Twisting is effective in reducing pickup interference from three-phase power wiring and in three-wire circuits, such as those used in servo wiring. When a twisted pair is being used, each lead of the pair should be tested to determine the lead combination that produces the least pickup voltage.

c. Multiconductor shielded cables are generally of two types. One type contains a large bundle of individually shielded single-conductor lines used to complete a number of low-level circuits. For the individual shields within this cable type to provide maximum effectiveness, it is essential that each be individually insulated. The other cable type contains a single shield over a large bundle of insulated wires; this cable type is generally used for transmission of control signals. The shielding primarily reduces crosstalk transfer from the shielded control cable into adjacent signal cables.

d. An unshielded cable may be effectively changed into a shielded cable by routing it through continuous metallic conduit. This routing is frequently done in power distribution systems. For economical reasons, the conduit is generally of aluminum; but from the standpoint of shielding high-level power circuits, galvanized steel conduit is more effective at power frequencies.

e. The proper installation of cables is essential if interference difficulties are to be avoided. Assuming proper grounding techniques have been employed, the following are suggested as guidelines for good signal cable practice:

1) Shields should not be used for signal return circuits

2) All signal circuits, including signal ground returns, should be individually shielded and have insulating sleeves or coverings over the shields. Balanced signal circuits should use twisted pair or a balanced coaxial line with a common shield. Where multiconductor twisted pair cables, that have individual shields as well as a common shield are used, all shields should be insulated from one another within the cable

3) Coaxial cables should, in all cases, be terminated in their characteristic impedance

4) On shielded cables in harnesses, where a common shield ground must be utilized, a clamp or heavy conductor (halo) should be used to ground all shields to the connector body. This should be done in addition to connecting the shields to ground through one or more connector pins

5) Coaxial cables carrying high-level energy should not be bundled with unshielded cables or shielded cables carrying low-level signals. Although the characteristic impedance of a cable or signal circuit should normally be quite low, the shield-circuit impedance may become appreciable if the shield becomes open ended, or electrically long. This reduces shielding effectiveness

6) Shields should be grounded on both sides of a connector to avoid discontinuity. If this is not possible, the shield should be carried across the connector through a connector pin

7) Grounding a number of conductor shields by means of a single wire to a connector ground pin should be avoided, particularly if the shield-to-connector or connector-to-ground lead length exceeds one inch, or where different circuits that may interact are involved. Such a ground lead is a common-impedance element across which interference voltages can be developed and transferred from one circuit to another

Great care must be taken at connectors if impedance characteristics and shielding integrity are to be maintained. A shielding shell should be used to shield the individual pins of a connector; a well-designed connector has a shielding shell enclosing its connecting points (fig. 2-86). The shell of multipin connectors should be connected to the shield. Coaxial lines should terminate in shielded pins. The use of pigtail connections for coaxial lines is undesirable since it permits rf leakage.

### 2-30. Cable Shield Grounding

a. Each shield circuit should be carried individually; each should be electrically continuous and grounded at both ends. In the case of long cable shield runs, bonding of shields at intermediate terminals, or locations, is advisable to reduce impedance of the shields to ground, thereby rendering the shielded circuits less susceptible to radiated or induced interference. Individual shields should not be electrically joined together so that one shield carries the rf currents of another. To obtain maximum rf shielding from shielded wires or coaxial lines, it is necessary to bond them effectively to the ground plane. For a low-impedance rf connection, the shortest length of connecting strap or jumper that is mechanically practical should be used, and the bonding procedures outlined in section 3 of this chapter should be followed. If coaxial cables are used to transmit rf signals, they should be grounded at both the sending and receiving ends. Normal coaxial connectors are adequate for this purpose; pigtail connections should be avoided. In applications where twisted-pair cables are used, the shield should be grounded at each end, and the circuit return path should be floating (single-point grounding). Bonding and grounding techniques employed should comply with applicable MIL Standards of Good Installation Practice.

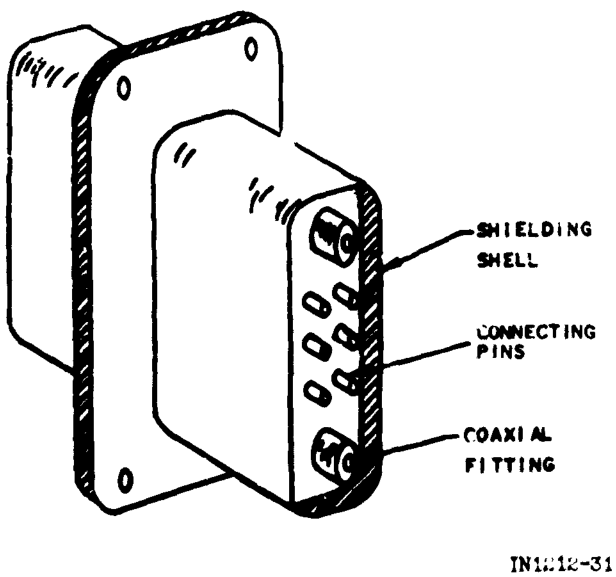



Figure 2-86. Connector With Shielding Shell Enclosure For Connecting Points

b. Both multipoint and single-point ground systems offer singular design features. When electronic and electrical systems are distributed over a large area, multipoint shield grounding is superior for interference control. Further, the multipoint approach allows short ground connections, provides a low impedance ground plane between system intertie points, permits a low-impedance ground return circuit, and improves the effectiveness of filter installations. While multiple-ground circuits are recommended for rf applications, there are some circumstances, primarily in low-frequency, low-level work with audio or servo amplifiers, in which single-point grounding is necessary. When a shielded cable in a sensitive circuit is grounded at both ends for the return circuit, power-frequency currents in the ground plane can induce audio-frequency interference. Therefore, single-point grounding may be the best approach where large ac currents, flowing in the ground plane, may couple into extremely sensitive low-frequency circuits (fig. 2-87). To provide extra protection, a shielded, twisted pair should be used (fig. 2-88). The shield should be grounded at both ends; the signal return lead, however, only at one end. Because of multiple grounding of the shield, magnetic fields may be coupled into the shield by conduction or induction. The twisted leads reduce magnetic susceptibility because of field cancellation.

c. Serious interference problems arise when shielded wires or coaxial cables are not properly terminated at the connector. It is important that the connector be properly grounded. The direct bond for this ground can be achieved by maintaining clean metal-to-metal contact between the connector and equipment housing. In those cases where a large number of individual shields from shielded wires must be connected to ground, it is recommended that the halo technique be used. The exposed unshielded leads should be as short as physically possible to reduce electrical coupling between conductors. Interference is caused when a shielded cable is run into a completely sealed box, but is grounded internally. The correct way to install a shielded rf cable is to run the shield well inside the connector and bond it around the connector shell. The arrows on figure 2-89 show the path that any signal or interference, that is picked up on the outer surface of the shielding, must follow to return to ground. The currents around the loop generate a field in the enclosed box, as do coupling loops used with resonant cavities. Figure 2-90 illustrates the correct method of introducing

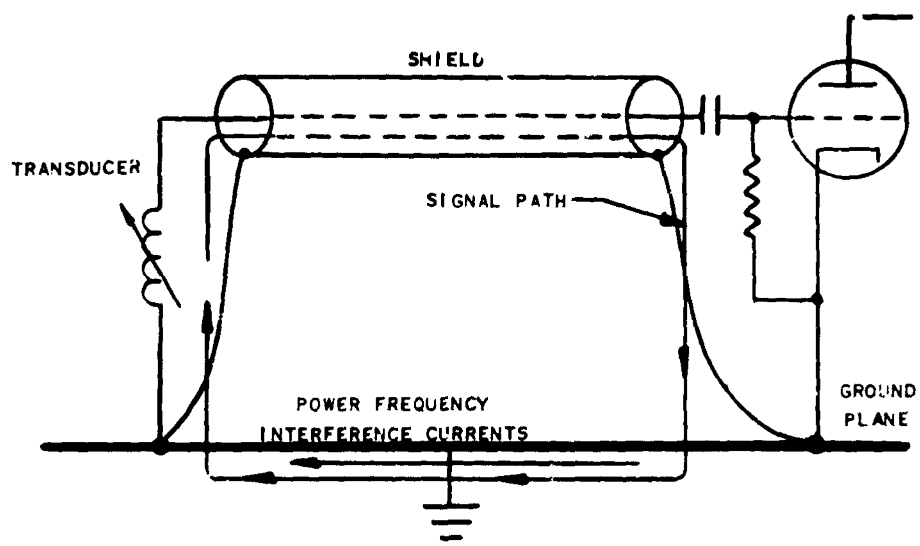



Figure 2-87. Power Frequencies Coupled Into Low-Level Circuit Using Multipoint Shield Grounds

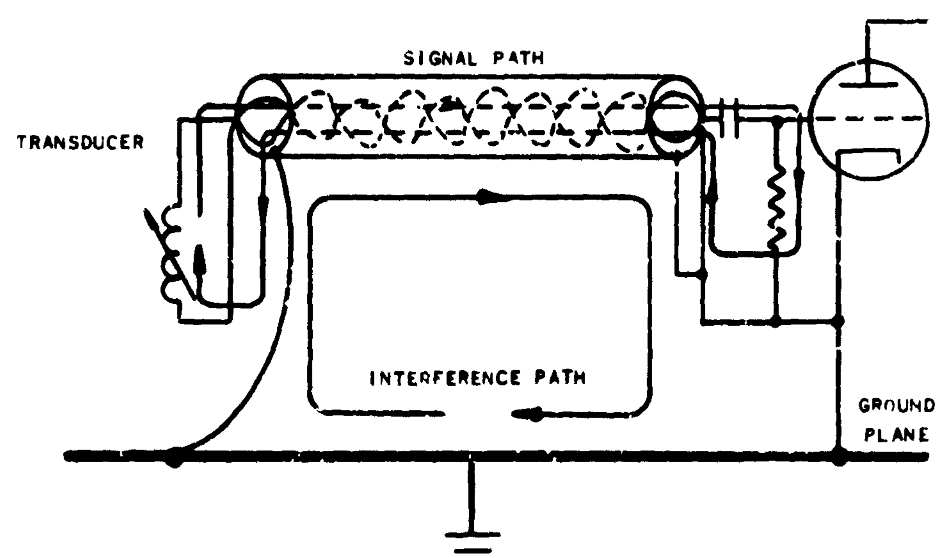



Figure 2-88. Reduction of Signal and Power Frequency Coupling by Shielded, Twisted-Pair Cable

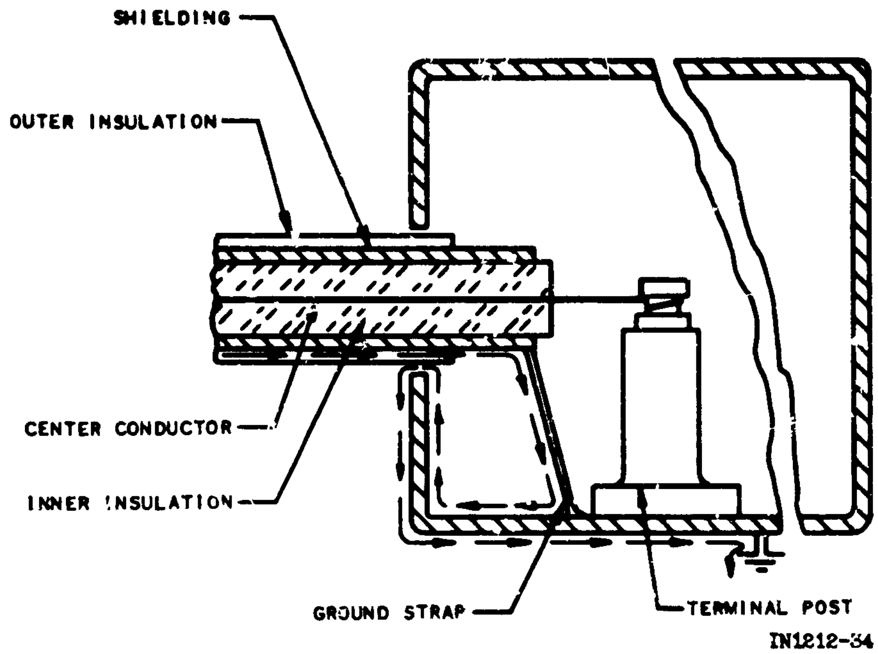



Figure 2-89. Incorrect Method of Introducing Shielded Cable

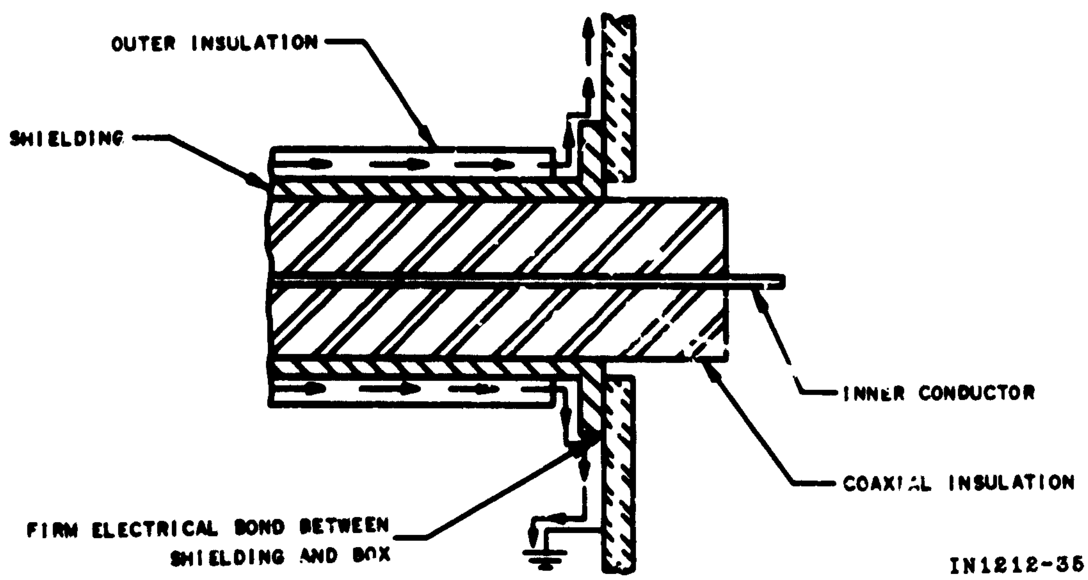



Figure 2-90. Correct Method of Introducing Shielded Cable

shielded cables into a box where shielding must be maintained. Interference currents may be carried when a shielded rf cable entering an enclosure has its shield stripped back to form a grounding pigtail. Such pigtails should therefore be avoided. If it is absolutely necessary to use a pigtail, it should be kept as short as possible and soldered to provide a ground without breaking the shield. The pigtail should also maintain continuity of the shield (through a pin in the connector) to a continuation of the shield inside the enclosure. The cable rf shield is a part of the complete shielding enclosure and should have no openings (fig. 2-91).

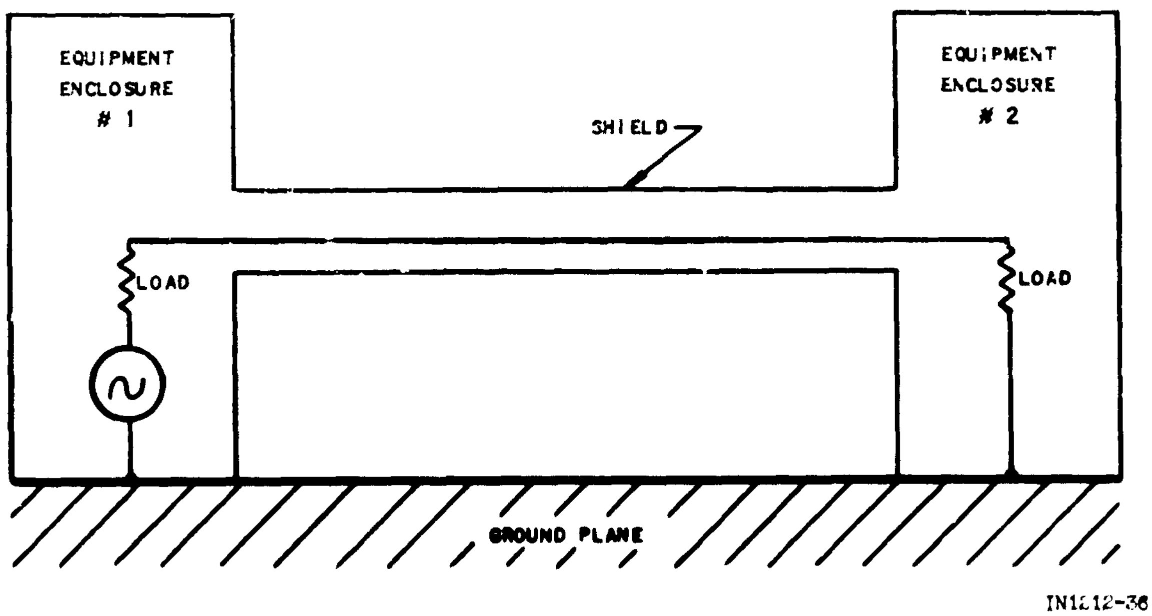



Figure 2-91. Continuous Equipment Enclosure Shield

d. During initial design stages, consideration should be given to proper location of equipment and wiring to minimize interference coupling between transmission paths. Many interference problems are eliminated by suitable physical arrangement of individual components and equipment. Sensitive equipment should be kept as far as possible from units that may be sources of electrical interference. Cabinet panels or partitions should be used to separate or shield these components.

Power leads, control wiring, and other cabling to sensitive equipment should not be close to any interference source, or leads that may be carrying interference because of the inductive coupling that can exist between wires. The bundling of several leads together should also be avoided because of the possibility of interference coupling. If it is necessary for sensitive signal leads to pass near interference-carrying leads, relative orientation should be as nearly at right angles as possible to minimize the coupling effect. The distribution of power through multiple lines, from a primary power source to the components of a piece of equipment, is recommended to reduce component interaction. The signal circuits should be separated from ac power circuits and any other circuits that can transfer interference to them.

e. The use of shielded hookup wire for susceptible leads inside a chassis helps to prevent interference signals from flowing to external leads where they can be radiated. Shielded hookup wire acts as a lossy transmission line at high frequencies and introduces considerable attenuation in addition to the bypassing effect. In using shielded wiring, the shield braid at the ends where connections have to be made should be pared back for minimum length to keep the shielding as complete as possible. The ends of the shielding should be connected directly to the chassis. The shielding should be bonded to the chassis at convenient points along the length of the lead. Leads that run side by side, or cross over each other, should have their shields bonded. When leads come to an outlet for external connections, the inner wires should be capacitor-bypassed to the chassis with the shortest possible leads. This bypassing helps to destroy resonances in the wiring and also helps to short-circuit any harmonics. The bypass capacitors should not be too large, usual values being approximately 500 micromicrofarads.

f. Electric plugs and receptacles are usually mounted on the front and/or rear of the equipment chassis, or on the mounting base. If electric receptacles are on the front of the case, the plugs should be separate units. Shield grounds should be made in accordance with figure 2-92 A. If electric plugs and receptacles are placed at the rear of the case, at least one unit should be securely attached to the case or chassis; the other should either be separate or securely attached to the mounting base. Shield grounds should be made in accordance with figure 2-92 B. Two poor methods of grounding cable shields are shown on figure 2-93. These methods are not recommended because their use permits interfering signals to enter the equipment. In cases where a common shield-ground must be employed, such as on multishielded cables, or in harnesses having a large number of individually shielded circuits, a clamp or bus should be used to ground all shields to the connector body, in addition to grounding them through one or more of the connector pins. This common ground should be avoided when the shield-to-connector or connector-to-ground lead length exceeds one inch, or when current circuits that may interact are involved. To prevent discontinuity of the shield because of possible disconnect at intermediate connectors, shields should be grounded to the structure on both sides of the connector. Where this is not possible, the ground should be carried across the connector, or through a conductor pin, to ensure continuity.

2-31. Cable-Shield Bonding

a. Shields should be terminated no further than 0.25 inch from the ends of the lines they are shielding. Bonding halos or interlacing straps should be used to terminate the shields and to minimize the impedance of the shield termination (figs. 2-94 and 2-95). Shields should be connected to the ground plane by 1.5 inches or less of 0.25- or 0.5-inch wide tin-plated copper strap. The halo technique is acceptable only when a few wires are involved. The interlacing strap method should be used for a common shield ground in multishielded cables or in harnesses that have a large number of individual shields. The interlacing strap should be at least 0.25-inch wide by 0.012-inch thick and be bonded securely to the connector (fig. 2-96).

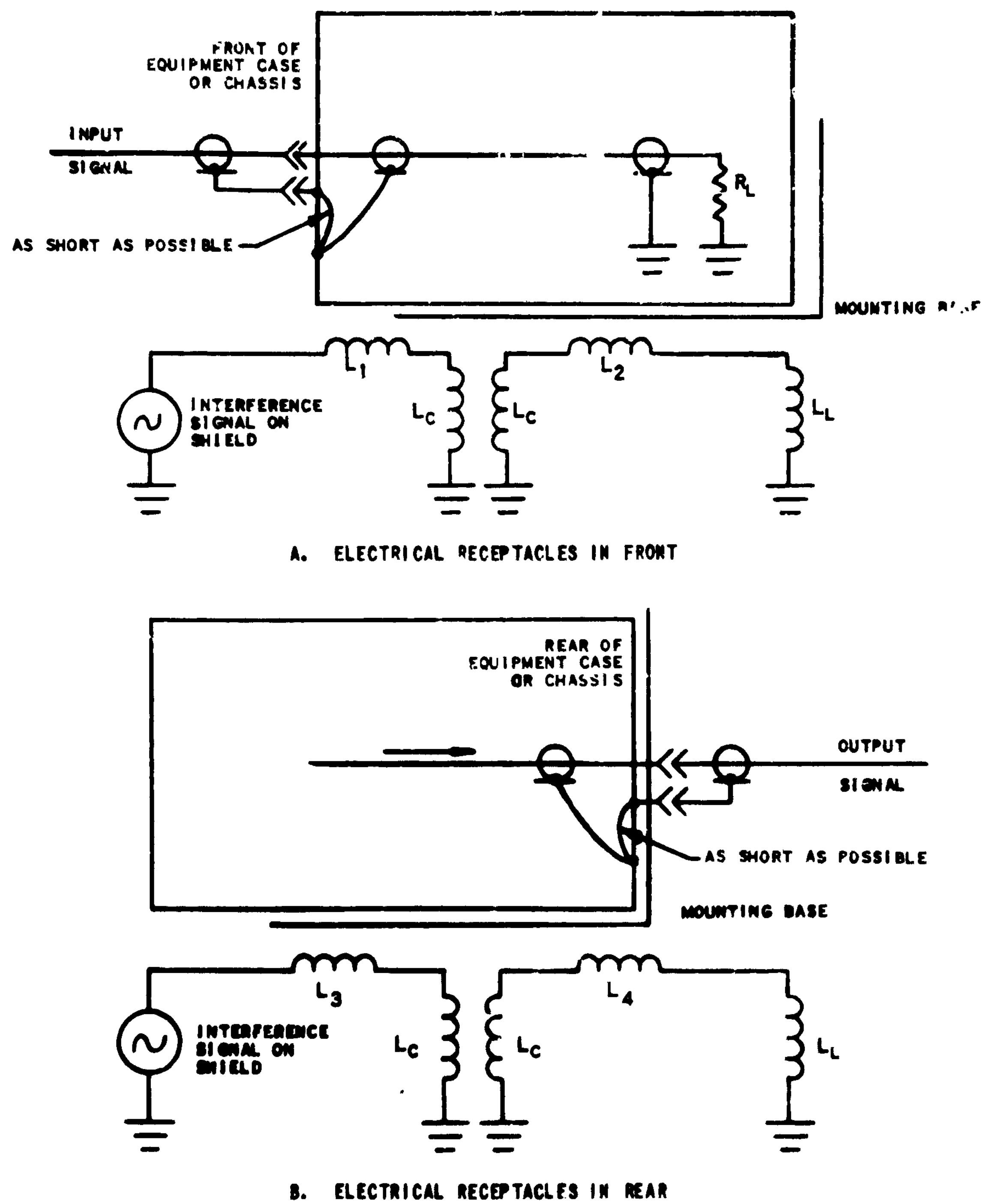



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Figure 2-92. Cable Shield Bonding

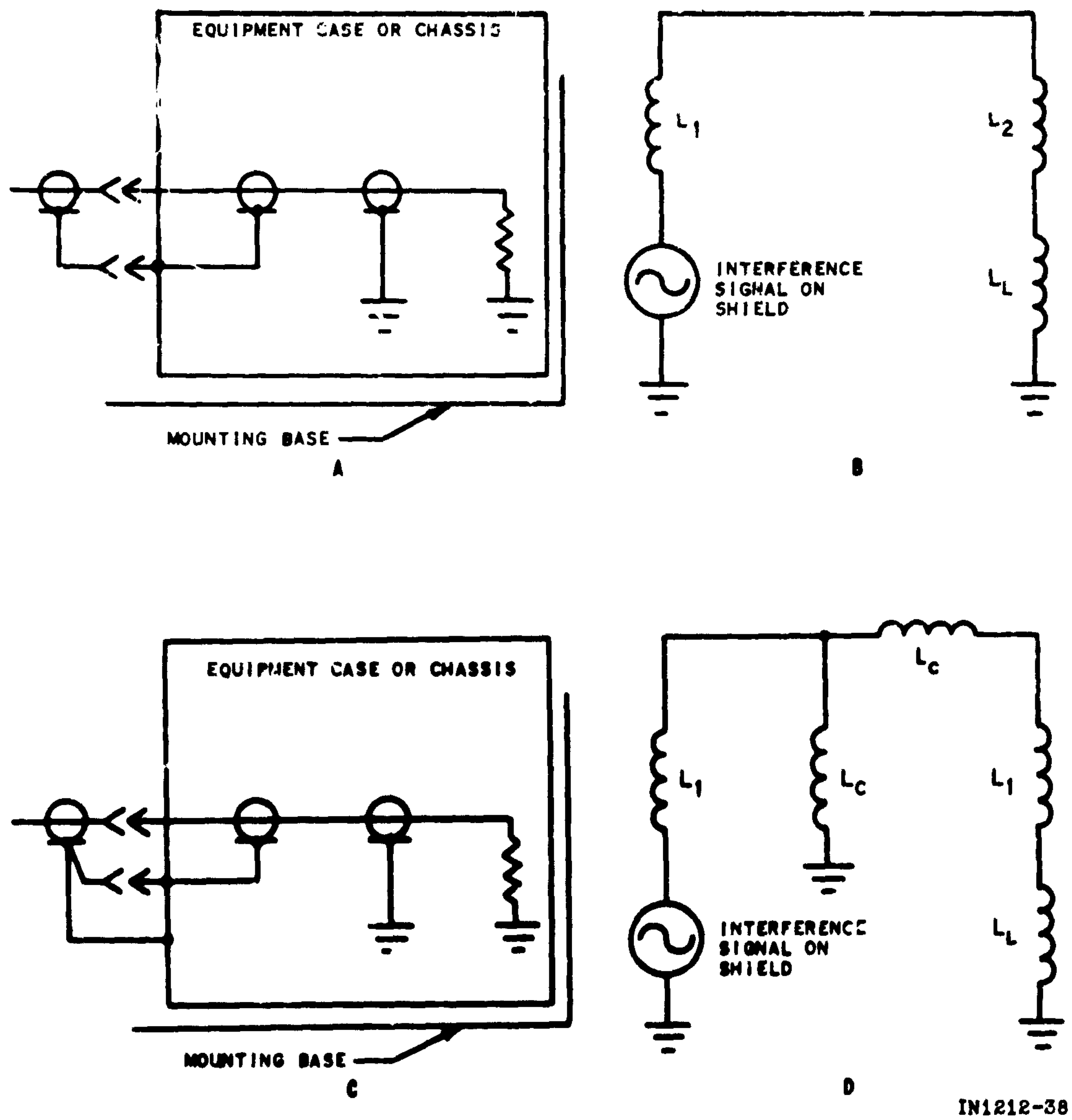
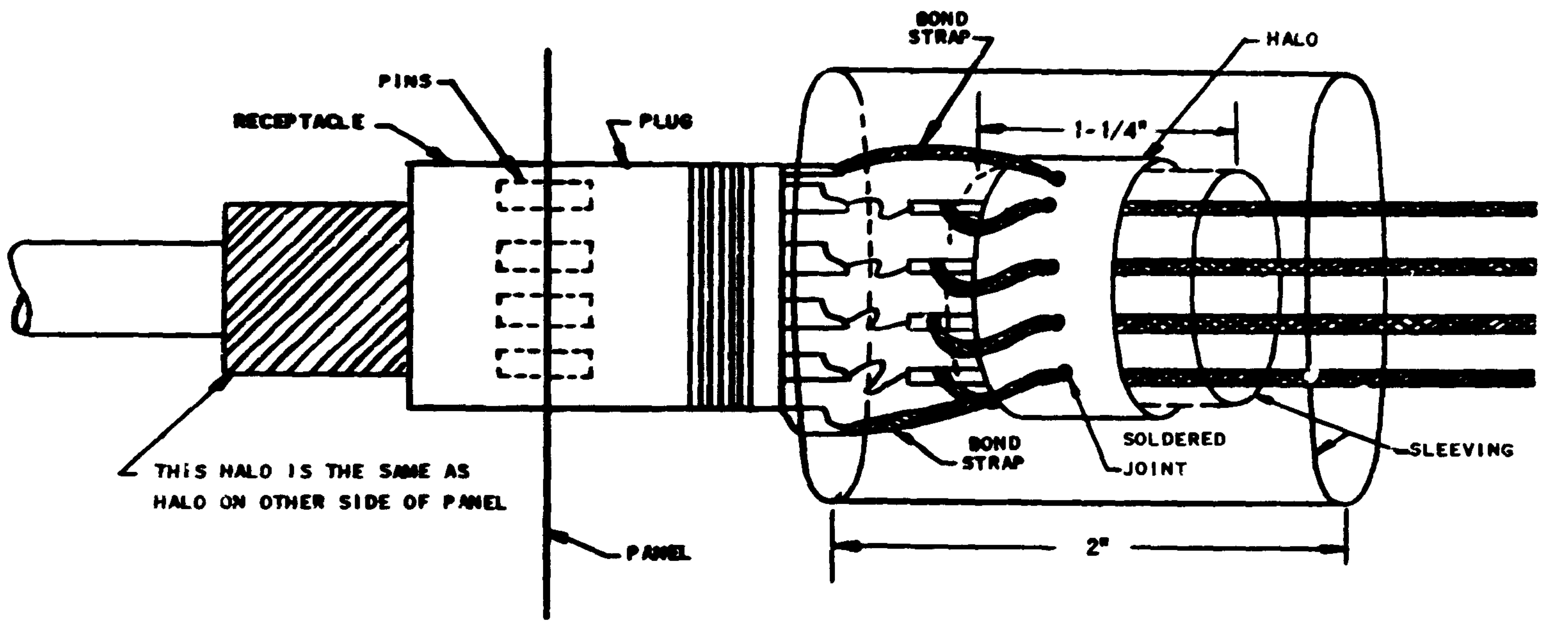



Figure 2-93. Poor Cable Bonding




NOTES: 1. BOND STRAP MAY BE CONNECTED AS SHOWN OR WITH 1/4 IN. BOND STRAP TIED TO STRUCTURE OR CONNECTOR BY MEANS OF EARED WASHER
2. HALO IS 1/4 TO 1/2 IN. WIDE

IN1212-40

Figure 2-94. Bonding Ring or Halo at Connector for Terminating Shields in a Harness

Figure 2-95. Typical Bonding Halo Applications

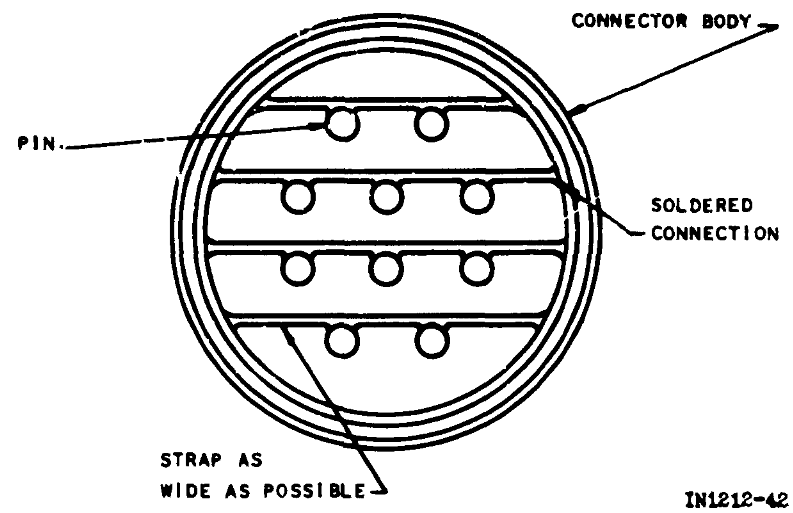



Figure 2-96. Interlacing Technique

b. Coaxial fittings should be kept tight at all times not only to provide a good impedance match, but to eliminate loose connections that might result in possible rectification of interference energy at the fittings. Shielding or bonding clamps that may be a part of the fittings should also be kept tight. Soldered fittings are recommended, particularly at terminations of shielding and braid.

c. The transfer of high-voltage pulsed energy requires considerable attenuation if the energy is to be confined within the desired circuitry without waveform distortion. The problem of interference from high-voltage lines can be eliminated by use of solid-copper transmission lines or triaxial cables; however, because of installation problems with the solid-copper lines, and difficulties encountered in long runs, triaxial cable is usually more convenient to use. In a typical installation, the center conductor is used to transmit the pulse; the first shield is used as the return; and the outer shield is utilized as a shield (fig. 2-97).

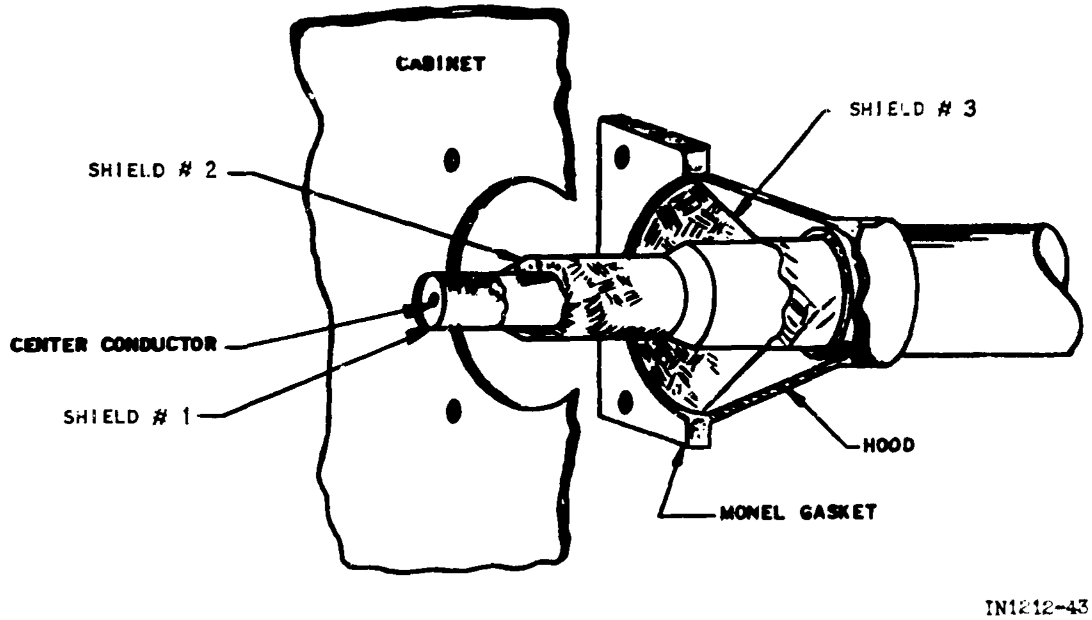



Figure 2-97. Triaxial Cable Application

2-32. Cable Routing

a. Crosstalk. Crosstalk occurs when signal information being carried by one cable or circuit is coupled into adjacent circuits. Crosstalk is a function of the type, frequency, level, and rate of change of a signal circuit source; level of the signal within the circuits experiencing interference; circuit impedances and match; and the degree of isolation between cable circuits. Isolation includes that afforded both by physical separation and by proper shielding. Crosstalk comprises signal information coupling induced by both magnetic and electric fields. Electric field coupling is a function of capacitance between conductors; magnetic field coupling is a function of mutual inductance between conductors. The inductive coupling between two wires depends upon the amount of flux linkage between them; the coupling between straight wires varies as the cosine of the angle between them. Thus, if it becomes necessary for signal leads to pass near interference carrying leads, they must always be as nearly at right angles to one another as possible. The amount of crosstalk present in a multiconductor cable is determined by the particular cable application. The application determines the geometry of the cable and such details as whether the conductors are to be parallel, paired, twisted, or shielded. In determining the geometry of the cable, proper physical grouping or routing of conductors can reduce crosstalk coupling through minimizing capacitive and inductive effects. In the routing of circuits, high-level, low-level, and pulse circuitry should be kept separate. This physical isolation, or separation, should be maintained within interunit cabling by routing each type of signal through a different cable. To reduce mutual coupling, cables should be separated from one another and routed away from interference sources and/or susceptible equipment. Physical separation between various circuits and cables is necessary because:

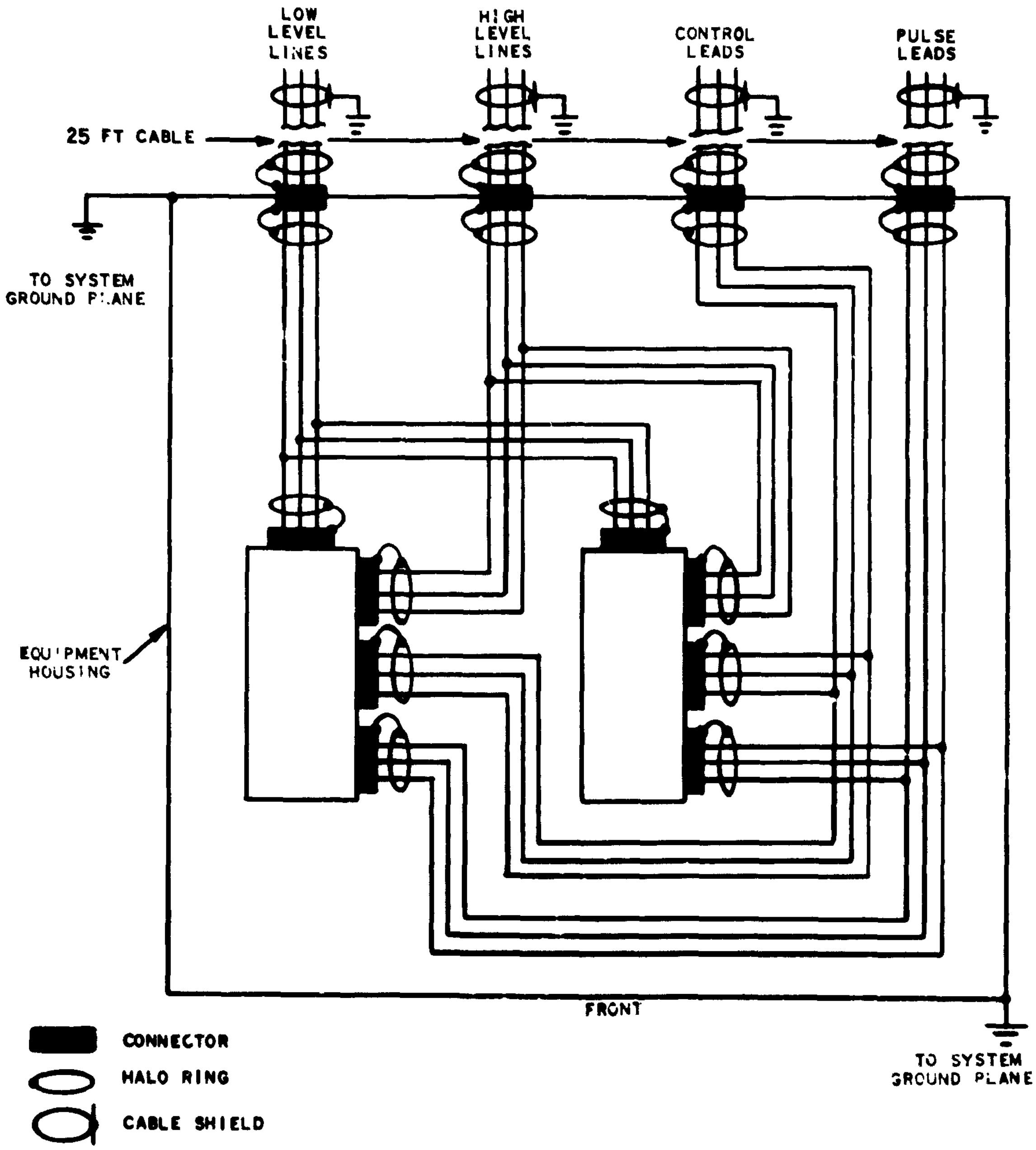
1) Low-level circuits are generally susceptible to crosstalk interference from high-level ac and pulse circuits

2) Pulse circuits radiate impulse-type interference and generate strong varying magnetic fields which easily transfer to other circuits

3) High-level dc circuits, when unshielded, can induce transients into adjacent circuits and cables

4) High-level ac circuits and cables become radiating interference sources for nearby circuits and cables

A typical installation employing many of the proper cabling methods discussed is shown on figure 2-98. In some installations, rows and tiers of cable trays are utilized to facilitate the desired physical cable separation. Detailed planning is required prior to installation to ensure that the best overall physical circuit and cable routing is achieved.




NOTES: 1. MODULES BONDED TO CABINET
2. ALL LINES INDIVIDUALLY SHIELDED AND INSULATED
3. LINES CROSS AT RIGHT ANGLES TO EACH OTHER TO EFFECT MAXIMUM INTERFERENCE REDUCTION

IN1212-44

Figure 2-98. Typical Cabling Methods

b. Circuit Coupling Effects. Circuits may be coupled by either mutual impedance or mutual admittance. These mutual elements may be resistances, capacitances, inductances, or any series or parallel combination of these elements. The degree of coupling can be greatly reduced by proper bonding and shielding. The mutual impedance of a common ground return, which can result from inadequate bonding, may become a major source of interference. Unless there is perfect shielding, there is always capacitance to metallic objects that will provide a return path for rf current through one of the circuits. Whenever there is a direct connection and return path between two circuits, a conduction current may flow between the circuits. The return path may be another metallic lead, a mutual capacitance, or a common ground return. The magnitude of the resulting current depends on the potential difference between points of entry and exit in the exciting circuit and the total loop impedance between these two points.

(1) Radiation Coupling. Radiation coupling will vary considerably with the path configuration and the frequency of the signal being transmitted. Generally, the degree of coupling increases as frequency increases. Such an increase may also result from increased antenna efficiency because any length of wire can act as an antenna and will be more efficient as its length approaches one-half wavelength. In a coaxial cable, holes in the braid may result in shunt coupling between the braid and center conductor of the cable; leakage may take place by wave penetration through the braid of the cable shield. Penetration occurs because of the finite resistivity of braid material and openings in the shielding. Leakage can also occur at the cable connector because of poor contact between the shield and the connector. It is difficult to shield against magnetic fields. Magnetic fields are a problem primarily at low frequencies, where the field appears as a low-impedance source. Little reflection from the shield

occurs, and reduction of the field must be accomplished by absorption in the shielding material. If high-permeability material is used, the shield will act as a low-impedance flux path and will divert the field. The shield should therefore be of high-enough permeability, or of sufficient thickness, to maintain the flux densities in it at levels below saturation. An insulating sleeve, or covering, over the shield of a cable prevents indiscriminate or intermittent contact between shields, or between a shield and a structural ground member. This contact might create interference from contact potentials or permit one shield to carry the shield currents of another. Precautions should be taken to prevent the exposed portion of a shield from intermittently contacting another uninsulated portion of shielded cable, as might occur during vibration. This can be done by rigidly tying down and lacing cables.

(2) Magnetic Coupling. The largest percentage of interference coupling encountered results from magnetic coupling -- energy transferred from one circuit to another through mutual inductance. Current flowing in a wire in one circuit produces flux which, in turn, induces voltage in a second circuit. The voltage induced depends upon the magnitude of the current producing the flux field, the length of the second circuit, the frequency of the interfering signal, and the distance between the two circuits (fig. 2-99). Circuit 1 on figure 2-99 represents the circuit causing the interference, while circuit 2 represents the circuit that has the interference voltage induced in it. $I_1$ is the current that is producing the magnetic field; $Z_{01}$ and $Z_{02}$ are the source impedances; $Z_1$ and $Z_2$ represent load impedances; $e_0$ is the interfering signal; and $L_M$ is the mutual inductance between the circuits. This mutual inductance is directly proportional to the length for which

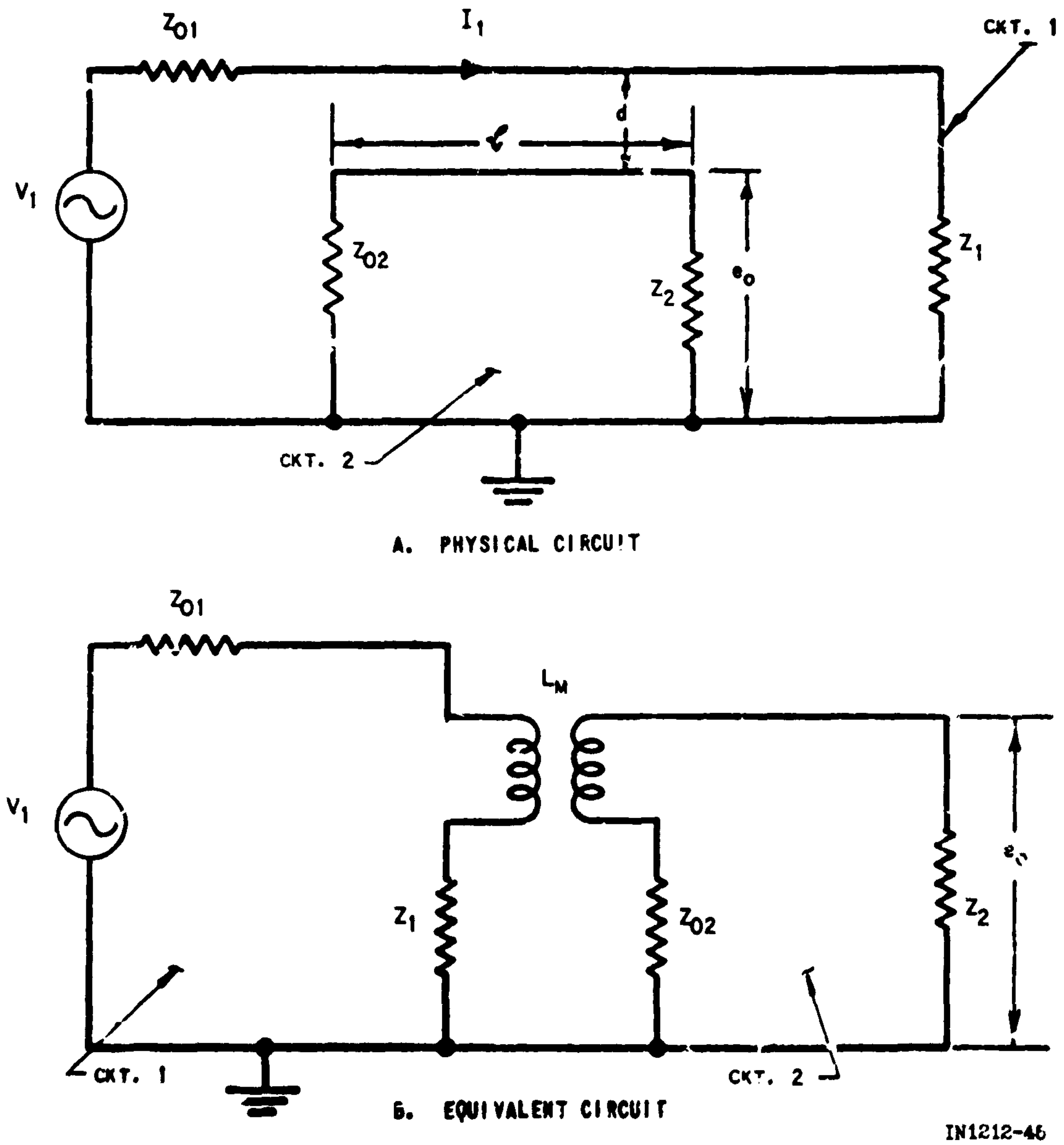



Figure 2-99. Magnetic Coupling

the circuits are coincident ($\ell$), and inversely proportional to the spacing between them (d). The voltage induced in the secondary of the equivalent transformer is proportional to the frequency, the mutual inductance, and the current in the source circuit. The interference voltage is the induced voltage multiplied by the ratio between the sensitive circuit load and the total impedance of the secondary:

$$e_0 = K f L_M I_1 \frac{Z_2}{Z_2 + Z_{02}} \qquad (2\text{-}53)$$

Any change in circuit parameters or configuration that changes the flux linkages changes the voltage induced. The physical separation of the circuits reduces voltage by lowering the flux density in the pickup loop. In addition, circuit configurations can be employed in which equal and opposite self-cancelling voltages are induced in the circuit. At frequencies below 5 kc, a twisted-pair will provide over 20 db of magnetic coupling reduction, while copper braid shielding will provide practically none. Conventional copper braid shielding for containment of magnetic fields becomes more effective as the frequency is increased above 5 kc. Ferrous shielding increases the shielding effect below 5 kc. For effective magnetic decoupling throughout the spectrum, twisted-pair conductors enclosed by conventional copper braid shield are usually employed. The shielding effectiveness of copper braid is less than 10 db up to 20 kc; it increases to 40 db at 1 mc, and to 100 db at 40 mc. The effectiveness of twisted-pair wires depends, to a large extent, on the uniformness and tightness of the twist employed; unshielded twisted-pair of 18 turns/foot is more effective for reducing magnetic coupling than is shielded twisted-pair of 6 turns/foot. Table 2-33 gives minimum twists/foot for common conductors.

(3) Electrostatic Coupling. Electrostatic coupling occurs when a voltage in a circuit causes a current in another circuit. The mechanism of the coupling is through the mutual capacitance between the circuits (see fig. 2-100). Among the physical factors that determine the magnitude of the interference signal are the proximity of the wires, the length of their common run, their distance from the ground plane, the frequency of the source signal, and the ratio of the impedances in the two circuits. Although magnetic coupling does not take place when source current is not flowing, interference may still take place in the case of

electrostatic coupling; that is, the source-circuit load may be of infinite impedance and still cause interference.

TABLE 2-33. WIRE TWIST[a]

| Gauge Number | Two Conductor (twists/foot) | Three Conductor | Four Conductor | Six Conductor | Eight Conductor |
|---|---|---|---|---|---|
| 6 | 4 | 3 | 2 | 1 | 1 |
| 8 | 5 | 4 | 3 | 2 | 1.5 |
| 10 | 6 | 4 | 3 | 2.5 | 2 |
| 12 | 7 | 5 | 4 | 3 | 2 |
| 14 | 8 | 6 | 4 | 3 | 2.5 |
| 16 | 10 | 7 | 5 | 4 | 3 |
| 18 | 12 | 8 | 6 | 6 | 4 |
| 20 | 16 | 12 | 8 | 8 | 6 |
| 22 & above | 18 | 16 | 10 | 9 | 8 |

[a]This listing represents the minimum number of twists/foot for wire routing and distribution systems.

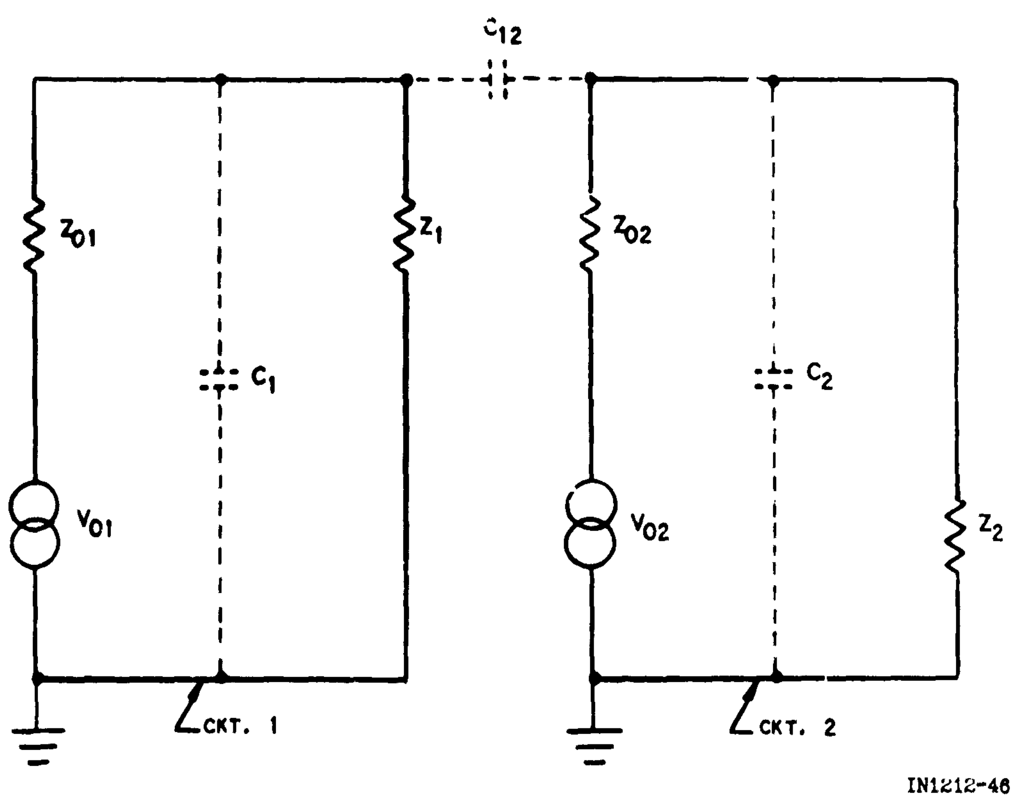



Figure 2-100. Capacitively-Coupled Circuits

c. Transient Response.

(1) Circuits, such as relay switching circuits, give rise to transients. A typical circuit of this type is shown on figure 2-101. Circuit 1 of figure 2-101 is activated by a step-voltage of magnitude $V_0$ at time t = 0. To simplify the transient response analysis, both circuits should be assumed to have the same capacitance to ground and the same parallel combination of load and source resistance. The following is the expression for the output voltage:

$$e_o(t) = \frac{1}{2} V_o \quad (R_p/R_{01}) \quad \left[ \epsilon^{-\frac{\gamma t}{\tau}} - \epsilon^{-\frac{t}{\tau}} \right] \qquad (2\text{-}54)$$

where: $R_p$ = parallel combination of load and source resistances of either circuit

$R_{01}$ = source resistance in circuit 1

$\tau = R_p C_1$

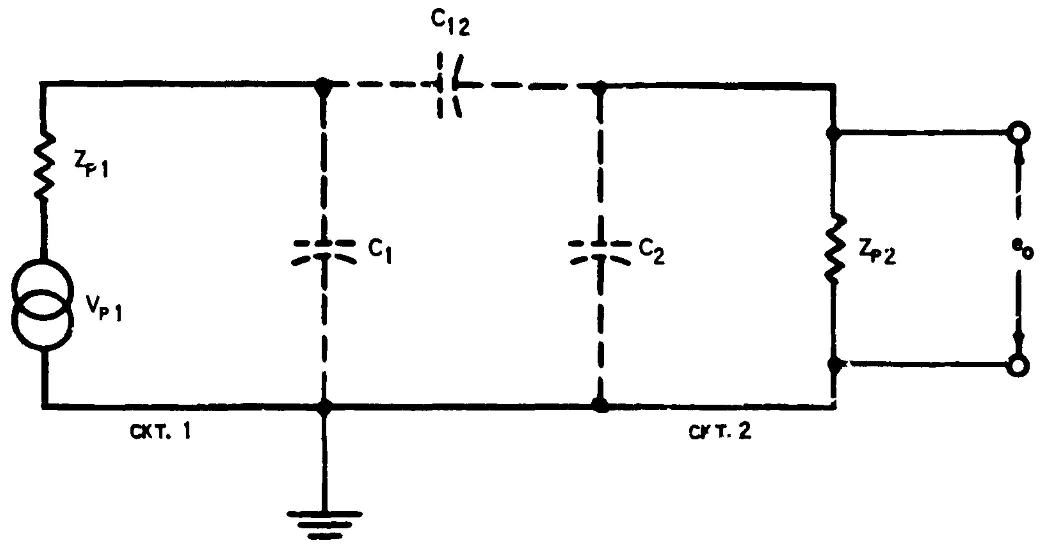
$\gamma = \frac{\alpha - 1}{\alpha + 1}$

$\alpha = 1 + \frac{C_1}{C_{12}}$

The waveform of this voltage is shown on figure 2-102. The maximum value of $e_0$ and the total energy (W) coupled into circuit 2 are given by the following expressions:

$$e_{max} = V_o \cdot \frac{R_p}{R_{01}} \cdot \frac{\gamma^{\alpha/2}}{\sqrt{\alpha^2 - 1}} \qquad (2\text{-}55)$$

$$W = \frac{1}{4} \frac{V_0^2}{R_2} \cdot \left(\frac{R_p}{R_{01}}\right)^2 \cdot \frac{\tau}{\alpha(\alpha - 1)} \qquad (2\text{-}56)$$




$$V_{P1} = V_{01}\,\frac{Z_1}{Z_1 + Z_{01}}\;;\quad Z_{P1} = \frac{Z_{01}Z_1}{Z_1 + Z_{01}}\;;\quad Z_{P2} = \frac{Z_{02}Z_2}{Z_2 + Z_{02}}$$

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Figure 2-101. Equivalent Circuit for Derivation of Coupled Signal

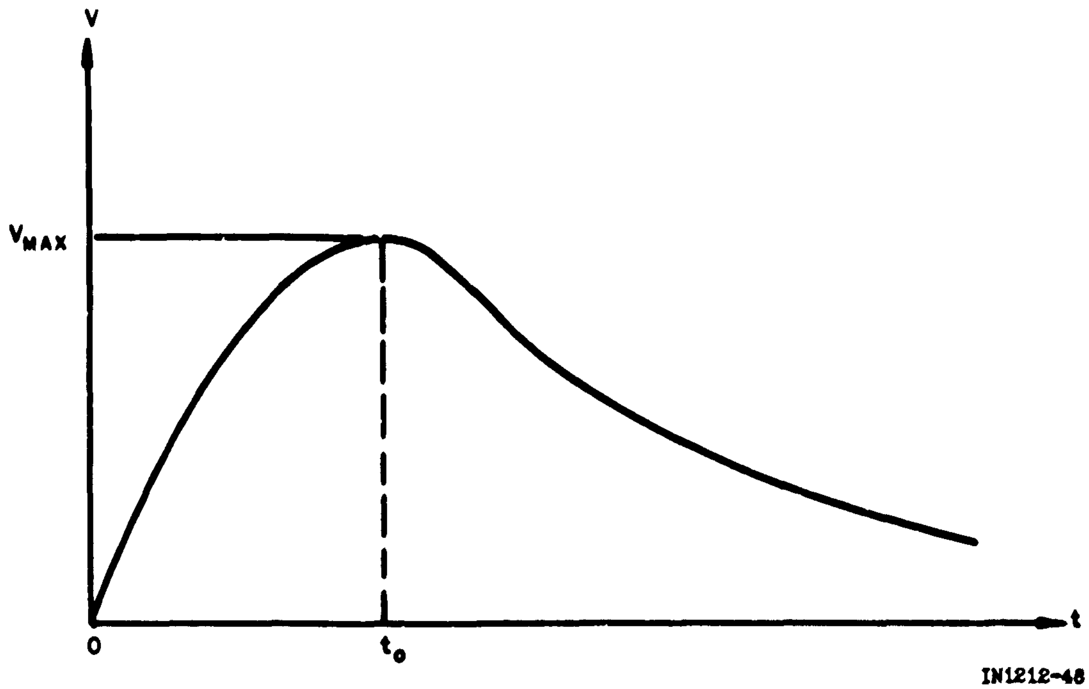


Figure 2-102. Coupling Signal Resulting from Step Voltage

Both parameters ($e_{max}$ and W) are sensitive functions of the coupling capacitance and the capacitance to ground of each circuit. In the absence of specific information about the nature of a circuit and its function, it is impossible to establish the relative importance of each of these parameters, and a general approach should be followed. Such a general approach is to compare pairs of circuits of different geometric configurations with respect to both parameters.

(2) A meaningful comparison between different circuit configurations requires the establishment of criteria for each. One basis on which to compare circuits is to maintain the same weight and volume occupied by the conductors for each. The coupling capacitance between two circuits and the capacitance to ground for each arise entirely from the proximity of the conductors in the flexible insulation.

(3) Cross-section views of cables of four circuits are shown on figure 2-103. The dimensions of the conductors and their spacing is the same for all four cases (fig. 2-104). Cases A and B differ only in the relative positions of the grounded conductors for the two circuits. Case C is the same as A, except that both sides of the insulation are covered with flexible aluminum or copper foil of approximately 0.001-inch thickness. The foil acts as a ground plane to provide electrostatic shielding and reduce capacitive coupling between adjacent circuits. Similarly, Case D is identical to B, except for the double shield. The peak voltage and the total energy transfer in all four of these cases is compared to a conventional shielded cable containing two twisted pairs. The capacitance in Cases C and D are the same; therefore, the results are the same and indicate that, of the cases considered, the double-shield configuration gives the best isolation between circuits. For purposes of comparison, the results are

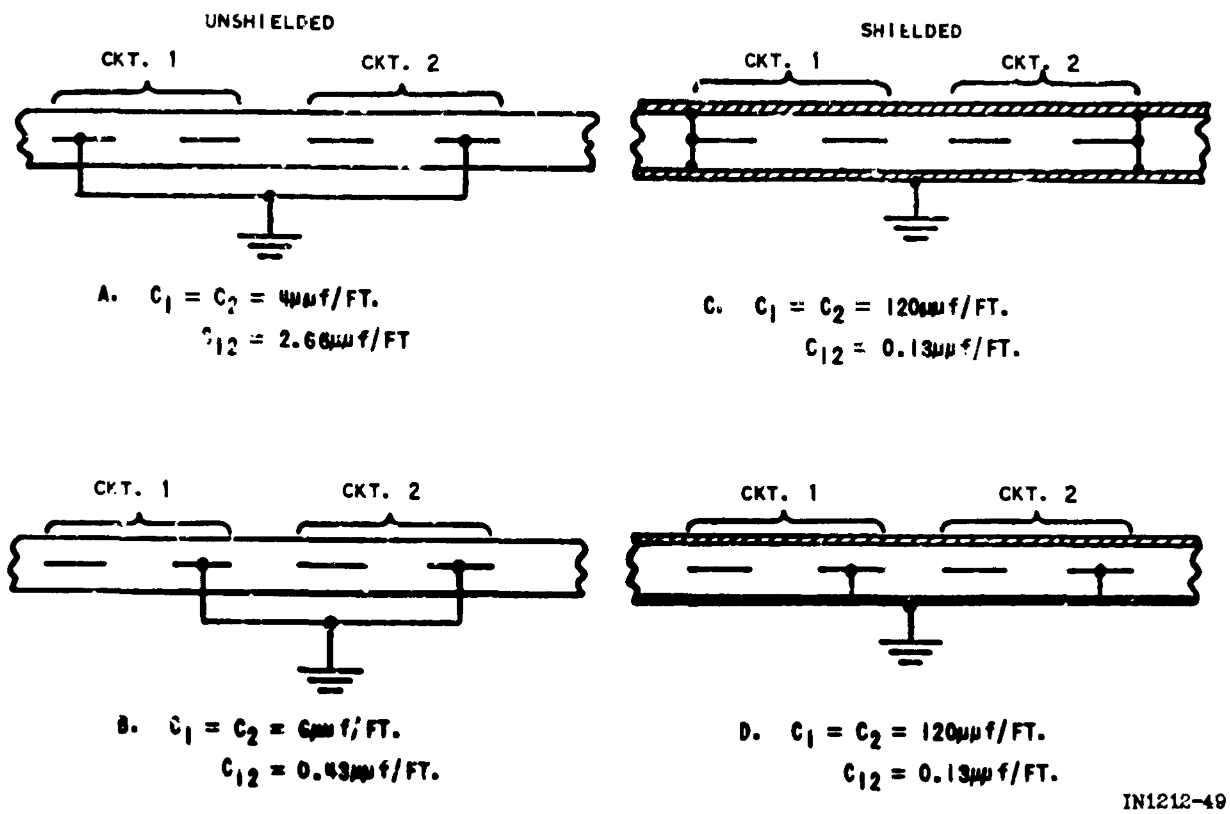


Figure 2-103. Cable Circuit Configurations.

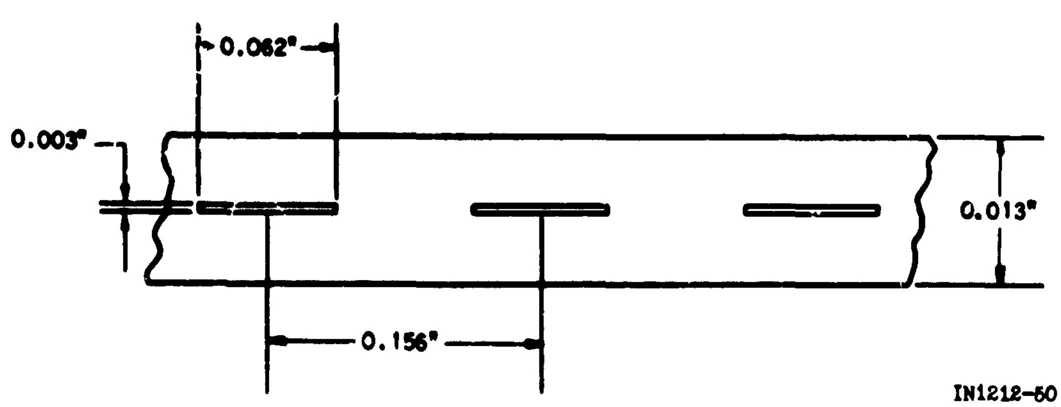


Figure 2-104. Dimensions of Flat-Conductor Flexible Cable

normalized for cases A and B with respect to the results for C and D. The results are:

| Case | $e_{max}$ | W |
|---|---|---|
| A | 380 | 7700 |
| B | 63 | 206 |
| C,D | 1 | 1 |
| Shielded cable | 133 | 1030 |

The listing shows the advantages gained by using the shielded configuration in dealing with circuits involving dc signals and with circuits involving switching that generates transients. It indicates that hazards due to voltage breakdown are more likely to occur in circuits generating impulse functions, and burnout due to coupled energy is more likely to occur in circuits involving dc switching. The comparisons shown in the listing serve to exemplify a possible means of predicting interference. It is possible to apply the same approach to other types of cable circuits; for example, printed boards and circular conductors. In addition, it is possible to make comparisons using such fixed parameters as cost, power handling capability and length of time that the interference signal is above a fixed threshold.

(4) To minimize capacitive crosstalk, the insulation should be of low dielectric constant, its thickness should be increased or, if possible, the two circuits should be routed in different layers (color groups) of multilayer cable.

d. Common-Impedance Cabling Elements. Interference signals originating in one unit of an electronic complex may be conducted and distributed to other units. A single signal, ground, or shield circuit may branch out to several distinct units, each located some distance from the other. This branching out may recur repeatedly to satisfy system intertie conditions. As this recurs, a network forms with common circuit elements. These common elements act as common impedances for currents originating in branches outside the common element, and voltages are developed across the common impedances. To reduce interference, it is necessary to eliminate as many common impedance circuit and cabling elements as possible. Such action requires careful design consideration within signal and ground circuits, but is readily accomplished within shield circuits. Common-impedance signal-circuit elements are best eliminated at the design stage by use of isolation elements such as individual amplifiers, cathode followers, transformers, or filters. Circuits that utilize a single-point grounding philosophy generally are not affected by this consideration as they usually do not consist of branching networks involving common impedance elements. Undesirable effects of common-impedance ground-circuit elements are effectively eliminated if an adequate ground plane, extending throughout the system, is utilized for the ground return circuits. While the ground plane itself is a common impedance element, its impedance must be low to preclude interference voltages developing across it of sufficient level to affect mutual ground circuits. Common impedance in shield elements can be effectively eliminated by utilizing the ground system and properly grounding all shield tie points. A ground loop is used to ground shields at each end, as is recommended for the reduction of interference. If an adequate ground plane exists between two pieces of equipment, or two locations at which the two ends of an individual cable shield are terminated, then the current through the shield will be greater than that between the two ground points and will not cause appreciable currents to be coupled into the signal circuit. The voltage that appears across the shield is effectively that which results from the current in the ground plane between the two points and the impedance of the ground plane.

e. Power Cabling. The three phases of each delta-connected transmission system, and the three phases and the neutral wire of each four-wire, wye-connected transmission system, should be twisted to form one cable. Twisting the wires together effectively cancels the electric and magnetic fields produced by the 120-degree phase-differential voltages and currents for either type of connection. Twisting the three-phase wires and the neutral wire in the four-wire, wye-connection effectively cancels the magnetic field produced by the in-phase third-harmonic currents that flow in each phase and add algebraically in the neutral. These third harmonic currents are generated when the iron cores of transformers or motors are driven to near saturation or operate in the nonlinear portion of the magnetization curve. The design of electrical equipment in which the iron cores are purposely driven into saturation is often undertaken where space and weight limitations are at a premium and the duty factor is low. Functional equipment performance degradation is a possiblility because of harmonic voltages and currents on the power lines. The wires should have as many twists per foot as is practical for their entire length. Where two wires are connected in parallel to form one phase of the three-phase system, a total of six wires is used and twisted to form a cable. Each conductor of a pair that represents one power phase of this six-wire cable should be selected diametrically opposite to its paralleling mate.

(1) Separation of Motor Loads From Signal Loads. A common source of interference is caused by commutation in dc motors and slip-ring friction in ac motors. Low-level signals are particularly susceptible to voltage transients generated by these sources. Separate power lines should be used to prevent conducted or radiated interference from entering into low-level signal lines.

(2) Separation of Utility Lines From Signal Loads. Utility lines may create undesirable paths for conducted interference signals and are quite capable of effectively radiating unwanted signals to nearby areas. Separate supply lines are needed to

preclude the entry of such interference into signal loads. A delta-to-wye hookup to provide 120 volts for single-phase signal loads and utilities is recommended. The lines supplying each load should be tied to a separate point in the junction box to isolate the circuits as much as possible. Individual grounds, returning to the earth ground, should be used in each.

(3) Placement of Conduit and Wireways. Low-level cables should be separated from other cables. Installation plans should identify high- and low-level cables for special routing and segregation. Cables used for pulse signals should be kept in separate wireways. The use of metal conduit is recommended in the installation of cables in which interfering signals cannot be tolerated. Separation of power supply lines and signal lines is necessary. Where cables and wireways are tied into junction boxes, separate boxes should be used for power and signal lines. If cables and wireways are routed into a single junction box, separate tie points and internal shielding between lines are needed. Of the two methods, the one using separate boxes is preferred. Open overhead wireways should be avoided, particularly in the vicinity of rf sources. Closed wireways and conduit, suitably grounded, are recommended.

(4) Balanced and Unbalanced Circuits. Balanced lines provide some degree of signal isolation. A circuit is considered balanced only when it fulfills all of the following conditions:

1) The wires constituting the circuit have equal and opposite current flow at any specified point

2) The circuit is free of primary ground plane grounding

3) Both sides of the circuit are of the same gage, length, and metallic material

4) Both wires are covered adequately with high dielectric insulation

5) Both wires are properly twisted with respect to each other

6) Metallic splices and terminations are solidly connected

(5) Signal Circuits and Pulse Leads. All signal circuits should be isolated from power or other circuits to the fullest extent practicable. The cabling or bundling of relatively low-level signal circuits with ac power circuits should be avoided. When the signal level is high, these circuits may be cabled together, if necessary, depending on the relative level between circuits. This expedient should be employed only after suitable tests have shown that no appreciable disturbance of the signal circuit occurs. Leads that intentionally carry pulse waveforms, similar to radar or pulse codes, should be run through separate connectors. When any pulse lead passes through a connector in a way that involves a discontinuity in the coaxial structure of the shielded lead (for example, when a pulse lead center-conductor and shield are attached to separate pins of a connector), an extremely low-impedance circuit should be provided for the ground lead of the shield in the connector. If this is not done, the entire shield along the lead may radiate or conduct interference. A suitable low-impedance connection can be obtained by the use of an extremely heavy pin, or by the use of several pins in parallel, preferably distributed circumferentially about the pin attached to the pulse lead center-conductor. The pulse lead and nearby susceptible leads should be shielded to prevent transfer to other leads. Additional protection can be provided by using rf filters to bypass unwanted interference components.

2-33. Waveguides

a. Where waveguides are used in electronic systems, interference can be generated by the creation of spurious modes of propagation as well as by arcing or by leakage at a waveguide junction. The generation of spurious modes should be resolved through proper guide excitation, dimensioning, and junction design. The installation of waveguides can also cause spurious modes because improper support in a vibration environment, particularly where flexible guide is used, can provide conditions whereby spurious modes can be generated. Prevention of interference from being generated as a result of arcing or leakage at a waveguide junction is predicated upon continuous conductive contact around the junction periphery. Good pressure sealing is required. A nonconductive, deformable rubber-like material should be used.

b. Because it is sometimes desirable to make a multipurpose seal for reasons of economy and availability, the same basic waveguide seal may be used in applications ranging from a waveguide feeding minute signals to a sensitive receiver, to a waveguide carrying the output of a megawatt transmitter. This universal seal should be designed for the high-power application and be capable of heat dissipation consistent with the waveguide section it adjoins.

c. At high frequencies, wavelengths diminish and short, physical discontinuities invite arc-overs. Tiny crevices and cracks in a seal become electrically significant. The seal must be designed so that no arcing will occur at operating frequencies within the field intensities that are expected to be encountered. Basic shielding theory assumes no discontinuities in the transmission line and no openings in the enclosure. Practical hardware will normally introduce discontinuities at a sealed joint. The design engineer should specify a seal that imposes a minimum of electrical discontinuity. The waveguide seal should provide a low impedance over the entire mating area, within the operating frequency range.

d. The standard choke flange, intended to prevent leakage, is a comparatively ineffective shield. This flange is effective, however, when modified by the addition of a woven-metal gasket in a separate outer groove to provide a low-impedance path between the flange faces. The choke flange has no broad bandpass and presents the possibility of resonance in variable frequency use. An effective seal should consist of a metal plate (matching the waveguide flange face) that has a pattern of raised contacts adjoining the waveguide opening. Adjacent to this electrical sealing area should be a contoured rubber gasket, molded into the metal plate, which deforms when the waveguide flange faces are closed down. The combined effect is a proven seal that accomplishes the dual purpose of leakage prevention and establishment of a low-impedance path between the flange faces. The pressure between the guide and the seal is a salient factor in controlling rf leakage as well as gas-pressure leakage. The loosening of even one bolt at a junction will produce a decided influence on rf leakage as well as gas pressure leakage.

## Section VI. EQUIPMENT MOUNTING

### 2-34. General

Situations arise that require shielding and bonding of equipment at the cabinet level. In the mounting of equipment, the cabinet functions as a bonding connection between the individual chassis and the ground plane, as well as an intermediary bonding intertie between the cable trays and the ground plane.

### 2-35. Rack Bonding

The equipment rack provides a convenient means of maintaining electrical continuity between such items as rack-mounted chassis, panels and the ground plane. It also serves as an electrical intertie for the cable trays. A typical equipment cabinet, with the necessary modifications to provide such bonding, is shown on figure 2-105. Bonding between the equipment chassis and the rack is achieved through the equipment front panel and the rack right-angle bracket. This bracket is grounded to the unistrut horizontal slide that is welded to the rack frame. The lower surfaces of the rack are treated with a conductive protective finish to facilitate bonding to the ground plane mat. The ground stud at the top of the rack is used to bond the cable tray to the rack structure, which is of welded construction. Figure 2-106 illustrates a typical bonding installation. The cable tray is bonded to the cable chute; the cable chute is bonded to the top of the cabinet; the cabinet is bonded to the flush-mounted grounding insert (which is welded to the ground grid); and the front panel of the equipment is bonded to the rack or cabinet front-panel mounting surface. Nonconductive finishes are removed from the equipment front panel before bonding. The joint between equipment and cabinet may have to serve a dual purpose: that of achieving a bond and that of preventing interference leakage from the cabinet if the joint is designed to provide shielding. If such shielding is a requirement, conductive gaskets should be used around the joint to ensure that the required metal-to-metal contact

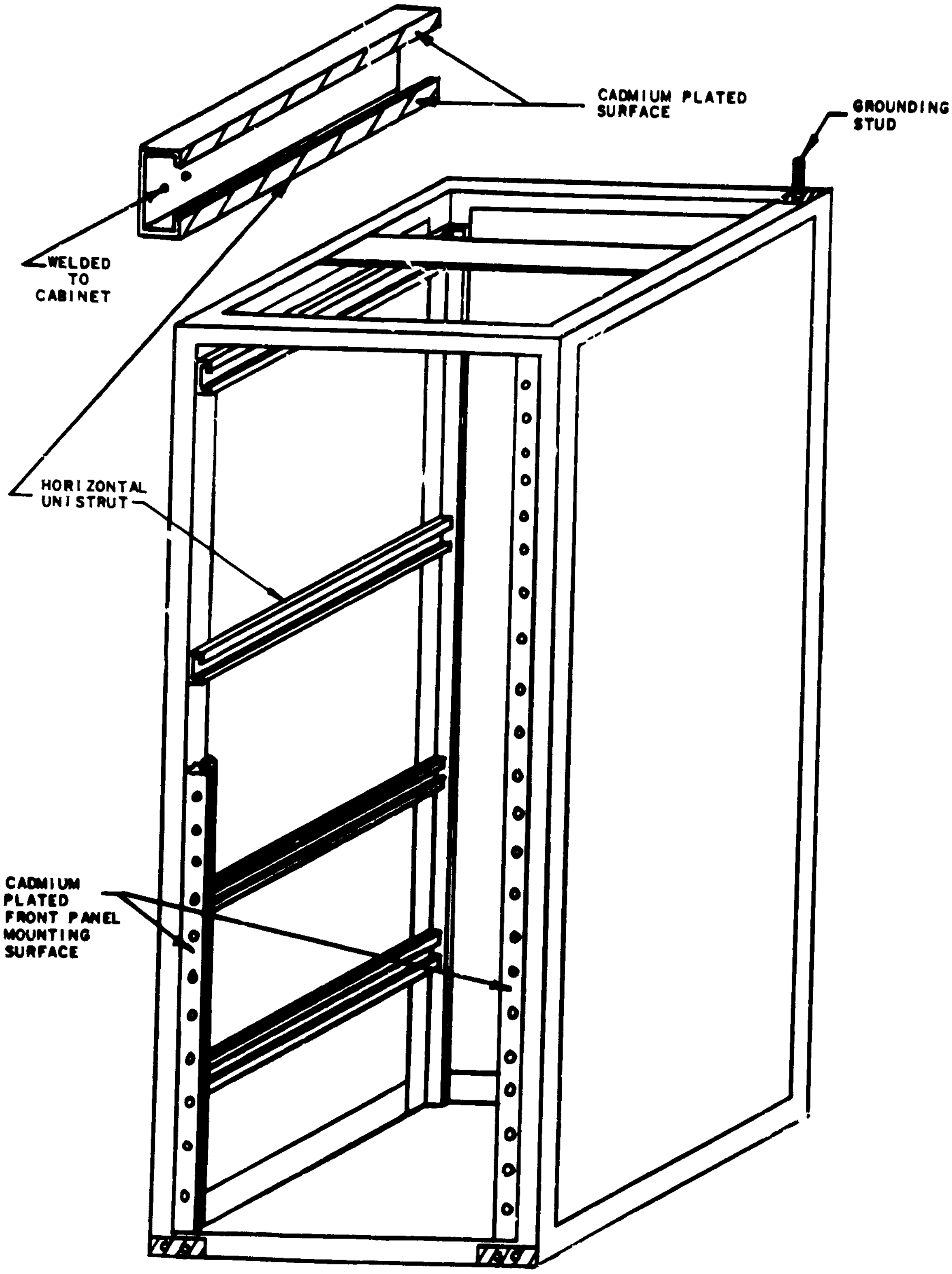


Figure 2-105. Cabinet Bonding Modifications

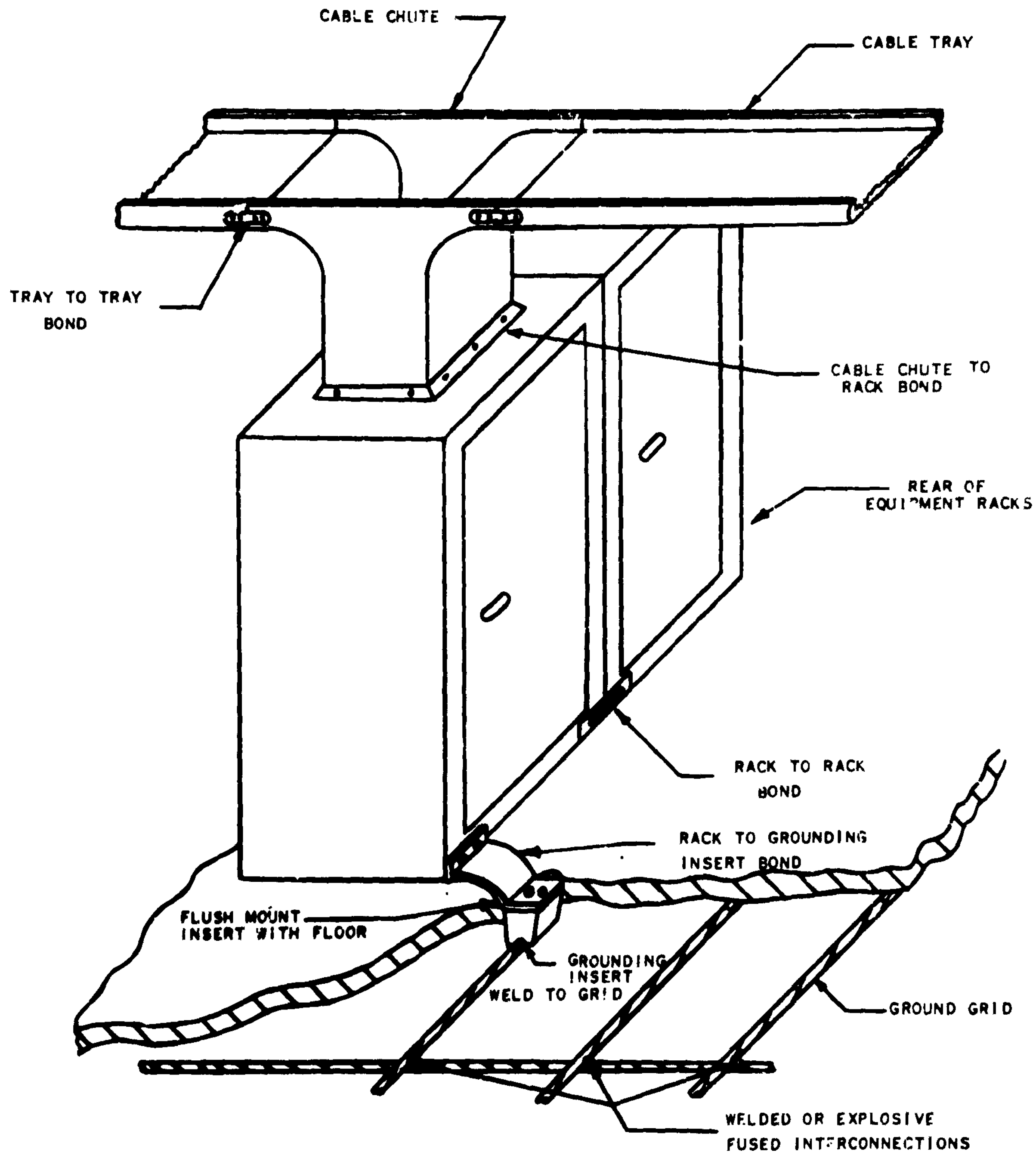


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Figure 2-106. Typical Cabinet Bonding Arrangements is obtained. If the equipment is in a shock-mounted tray, the tray should be bonded across its shock mounts to the rack structure. Connector mounting plates should use conductive gasketing to improve the chassis bonding. If chassis removal from the rack structure is required, a one-inch-wide braid with a vinyl sleeving should be used to bond the back of the chassis to the rack. The braid should be long enough to permit withdrawal of the chassis from the rack.

2-36. Cabinet Shielding and Equipment Location

a. It is often necessary to provide shielding at the cabinet level to contain interference generated by cabinet-mounted equipment or to protect such equipment from external electromagnetic fields, especially when unshielded off-the-shelf equipment must be integrated into an electronic complex. The interference characteristics of such equipment may cause electromagnetic compatibility problems unless external shielding is used. The procedures for establishing the cabinet shielding design are the same as for any type of enclosure. All joints should be designed to prevent leakage, and power circuits should be filtered. Cable entry into the enclosure is a critical area; a connector panel should be used for cable entry at the top of the cabinet. If entry is made through floor cable runs, it is possible to use an L-section in the bottom of the cabinet for access to the connectors.

b. Equipment location, such as the physical closeness of sensitive electronic circuits to potential interfering sources, can be a factor contributing to the interference problem. Where possible, equipment should be located to reduce the interference problem. In many cases, however, selective location and orientation criteria cannot be applied because functional relationships between pieces of equipment prevent them from being widely separated. Equipment housings should have a mating surface to fit into a readied well, or they can be connected together by means of beryllium-copper straps. These housings should be bonded to the ground system by beryllium-copper straps not less than one inch in width, which have a width-to-length

[redacted]

ratio of at least 1:5. At least 2 straps should be used for each equipment housing. The following location recommendations should be followed wherever possible:

1) Communications receivers and transmitters should be located so that their antenna lead-in wiring is as short as possible

2) Dynamotors, inverters, motor-alternators, and electric motors should be located in remote positions from receivers to prevent their case-radiated interference fields from coupling to the receiver lead-in wiring

3) All electrical machinery should be remote from openings in equipment structures to prevent interference fields from radiating directly to external antennas

4) Radar modulators and transmitters should be remote from communication receivers

5) Auxiliary power units should be remote from openings in the equipment so that ignition interference from the units cannot radiate to external antennas through the openings

6) Low-level circuitry should be remote from high-level circuitry